WIRELESS ENGINEER

THE JOURNAL OF RADIO RESEARCH & PROGRESS

AUGUST 1954

VOL. 31 No. 8 THREE SHILLINGS AND SIXPENCE

The PERFECT TEST TEAM

RADIO SHOW
Stand 61
EARLS COURT
Aug. 25—Sept. 4

The illustration depicts a set of modern "AVO" testgear being used to measure the "Q" of the secondary winding of the second I.F. transformer on a chassis of unknown characteristics—just one of many tests which can be performed by this combination of instruments.

A signal of predetermined frequency from the "AVO" Wide Range Signal Generator is being fed into the Electronic Test Unit, where it is amplified and fed to the secondary winding of the transformer. The Electronic Testmeter is connected across the tuned circuit under test and from the readings obtained and the controls of the Electronic Test Unit, the "Q" of the circuit can be determined.

The three instruments, shown as a team, cover a very wide field in measurement and form between them a complete set of laboratory testgear, ruggedly constructed to withstand hard usage.

ELECTRONIC TESTMETER

A 56-range instrument combining the sensitivity of a delicate galvanometer with the robustness and ease of handling of an ordinary multi-range meter. Consists basically of a highly stable D.C. Valve Millivoltmeter, free from mains variations and presenting negligible load on circuit under test. Switched to measure:—

D.C. Volts: 5mV to 10,000V.
D.C. Current: 0·5μA to 1 Amp.
A.C. Volts: ·1V to 2,500V R.M.S. up to 2 Mc/s.
·1V to 250V R.M.S. up to 200 Mc/s.
A.C. Power Output: 5mW to 5 Watts
Decibels: —10db to +20db. Zero level 50 mW.
Capacitance: ·0001μF to 50μF.
Resistance: ·2ohm to 10 Megohms
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ELECTRONIC TEST UNIT

For measuring small values of A.C. voltage, inductance, capacity, and "Q" at radio frequencies. Although designed primarily for use with "AVO" instruments, it can be used with any suitable Signal Generator/Valve Voltmeter combination.

As a Wide Range Amplifier, it is capable of an amplification factor of 40±2—3db between 30c/s and 20 Mc/s.

As a Capacity Meter, it covers measurements at radio frequency from ·5pF to 900pF in two distinctly calibrated ranges.

As an Inductance Meter, it gives direct measurements from ·5μH to 50mH in six ranges.

As a "Q" Meter, it indicates R.F. coil and condenser losses at frequencies up to 20 Mc/s.

Operates on 100-130v. and 200-260v. 50-60 c/s A.C. mains.

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WIDE RANGE SIGNAL GENERATOR

An instrument of wide range and accuracy for use with modern radio and television circuits.

Turret coil switching provides six frequency ranges covering 50 Kc/s to 80 Mc/s.

Range 1. 50 Kc/s—150 Kc/s
" 2. 150 Kc/s—500 Kc/s
" 3. 500 Kc/s—1·5 Mc/s
" 4. 1·5 Mc/s—5·5 Mc/s
" 5. 5·5 Mc/s—20 Mc/s
" 6. 20 Mc/s—80 Mc/s

Accuracy to within 1% of scale marking. Gives sensibly constant signal of good wave-form, modulated or unmodulated, over entire range. Minimum signal less than 1μV at 20 Mc/s and less than 3μV between 20 and 80 Mc/s. Gives calibrated output from 1μV to 50mV.

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And now a few words about service. Our unique Monthly Service covers primarily the despatch of "The General Radio Experimenter," a monthly house organ published by "G.R.". In this Journal is published all information concerning new "G.R." items, so that all readers are kept abreast of latest developments. In addition there often appear authoritative articles of great importance covering many aspects of laboratory measurements, as well as "production practice."

The "General Radio Experimenter" service is entirely free and there is no expressed or even implied obligation to make purchases. We do, however, request certain reasonable information from applicants for admission, these being: Surname, full Christian Names, full address of employers (if not self employed), and, if possible, type of work or chief interests. Please also state academic qualifications. Additions to our lists are entirely at our discretion.

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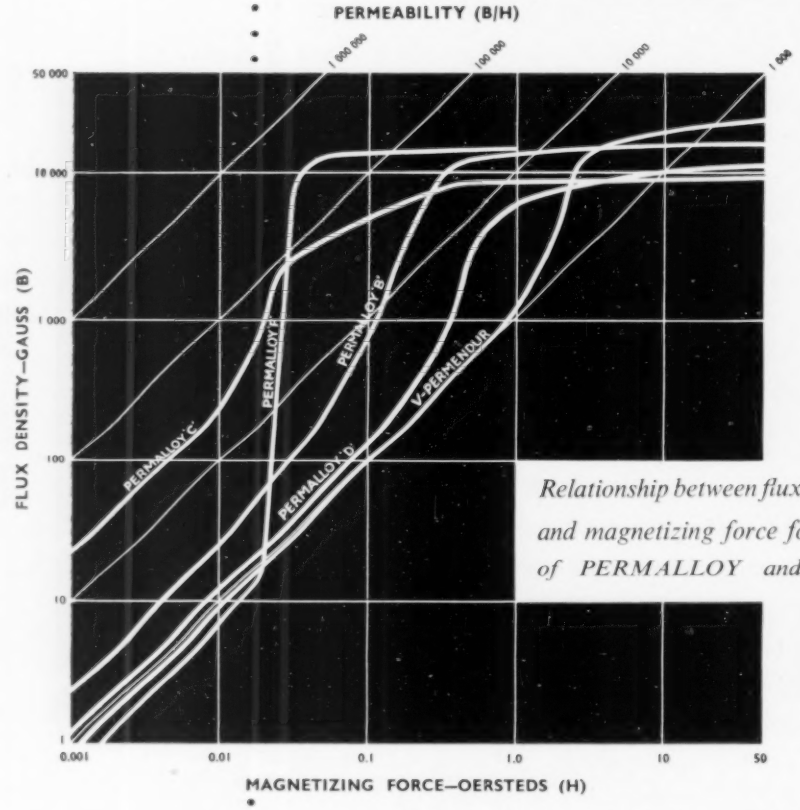

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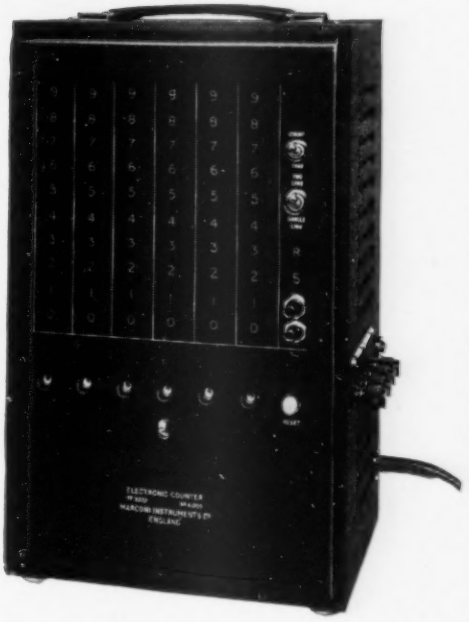

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The indication is given by illuminating the appropriate translucent figure in each of six vertical rows of 0 to 9 at the front of the instrument.

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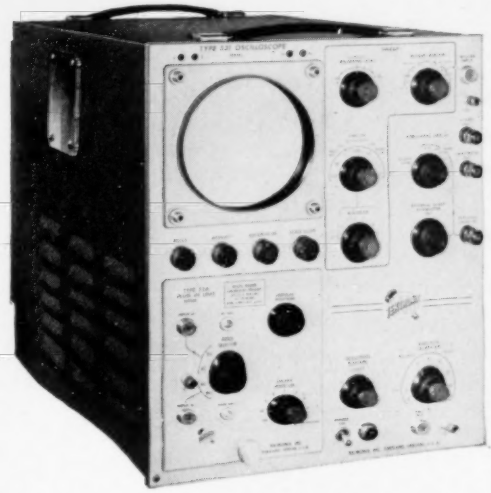

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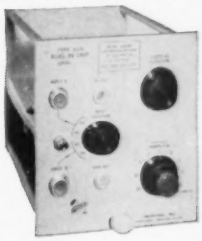
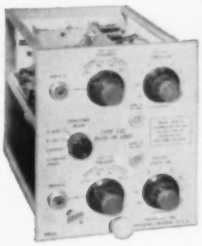
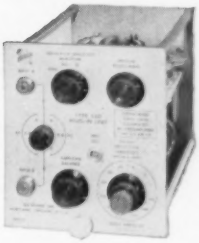


Type 53A Wide-Band DC Plug-In Preamplifier—dc to 10-mc passband, 0.035-μsec risetime. Sensitivity 0.05 v/cm to 50 v/cm, ac or dc, continuously variable, with nine calibrated steps from 0.05 v/cm to 20 v/cm. Two input connectors with 80-db isolation. Price $85.

Type 53B Wide-Band High-Gain Plug-In Preamplifier—same as the Type 53A with the addition of an ac-coupled input stage providing three additional calibrated sensitivity steps, 5 mv/cm, 10 mv/cm and 20 mv/cm. Passband 5 cycles to 9 mc, 0.04-μsec risetime. Two input connectors with 80-db isolation. Price $125.



Type 53C Dual-Trace Plug-In Preamplifier—two identical amplifier channels, each with dc to 8.5-mc passband, 0.04-μsec risetime, sensitivity 0.05 v/cm to 50 v/cm continuously variable with 9 calibrated steps from 0.05 v/cm to 20 v/cm. Electronic switching triggered by oscilloscope sweep, or free running at about 100 kc. Polarity reversal switches. Price $275.



Type 53D Differential High-Gain DC Plug-In Preamplifier—sensitivity 1 mv/cm at dc to 250 kc—with passband increasing to 750 kc at 50 mv/cm and lower. Sensitivity in calibrated steps—1 mv/cm to 50 v/cm, or continuously variable—1 mv/cm to 125 v/cm. Differential input. Price $145.

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FEATURES

Frequency range: 1-11,110c/s and 10-111,100c/s.
Frequency accuracy: ±0.2% or ±0.5c/s.
Hourly frequency stability: ±0.02% over most of range.
Maximum output:
2W into 8000 ohms above 20c/s.
50mW into 8000 ohms below 20c/s.
Harmonic content: 1% at 1W output.
Hum level: —80db relative to maximum output at 1000c/s.
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Weight: 83 lb. (38 kg).

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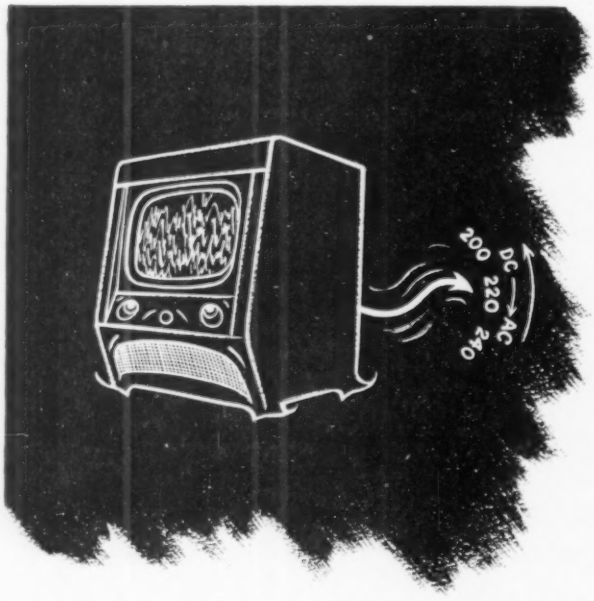

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THE TIMES and special supplements

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That is why the supplement for the radio industry published separately by THE TIMES on August 23rd will be so well worth reading. The articles, on varied subjects ranging from the radio show and current trends in set design to detailed news of the latest developments in components, will be informative and authoritative to a high degree. Reports on the B.B.C.'s plans for sound and television, a special article on the use of navigational aids in the air and on the sea, communications equipment, electronics as an aid to production, and many other articles of general interest are included.

This supplement of 32 pages published by THE TIMES reviews the radio industry fully and lucidly. Those who serve that industry, and the public in general, will find it stimulating and valuable reading.

THE TIMES

Radio and Television Supplement

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RADIO SHOW 1954

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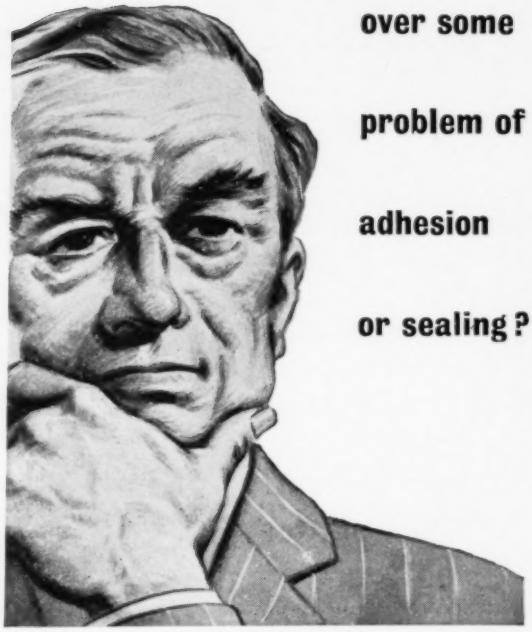

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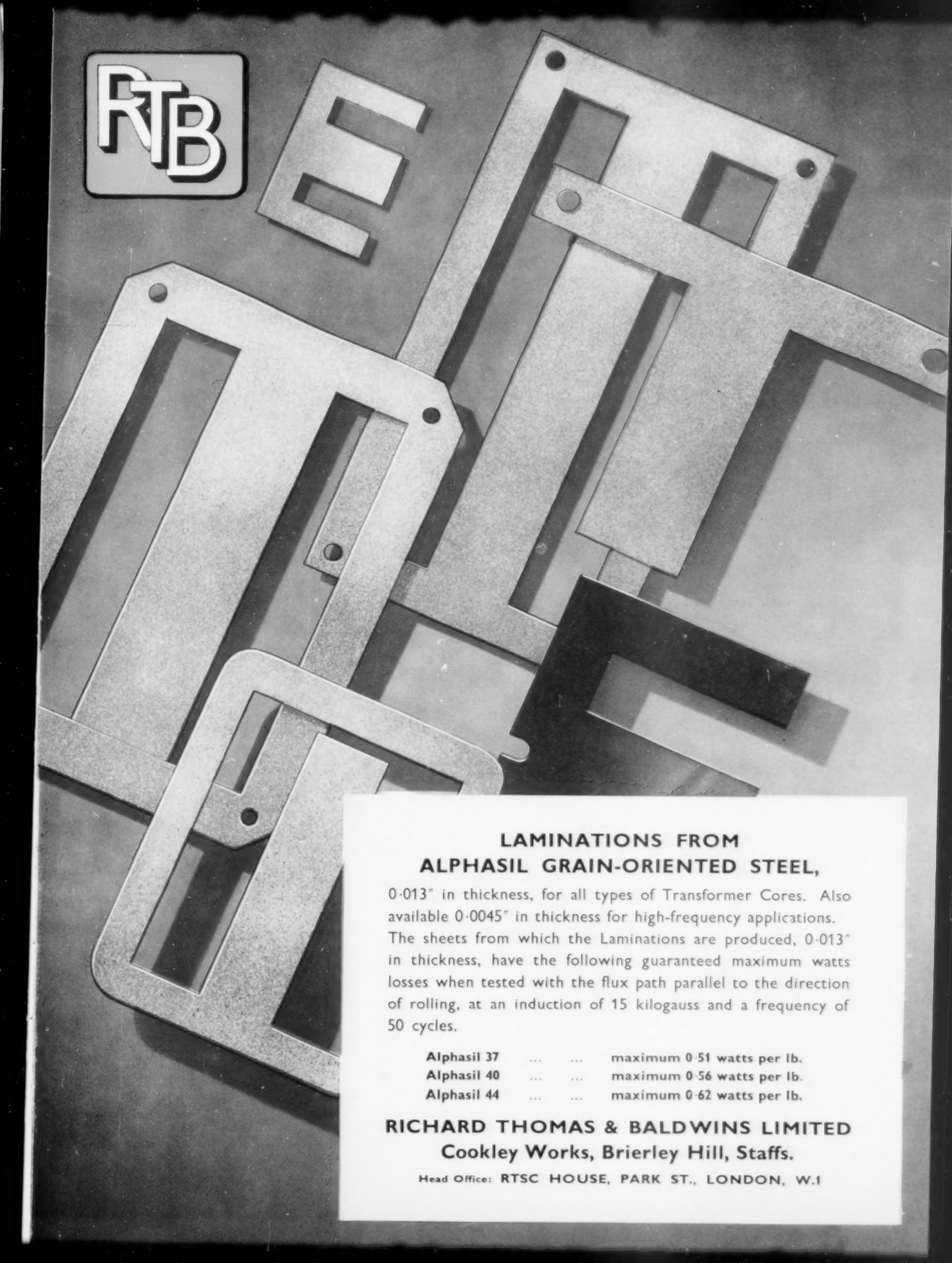

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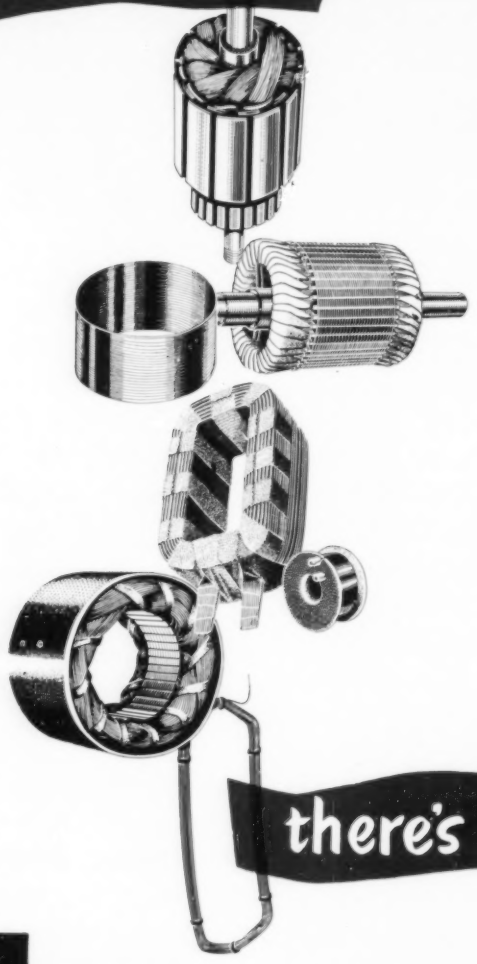

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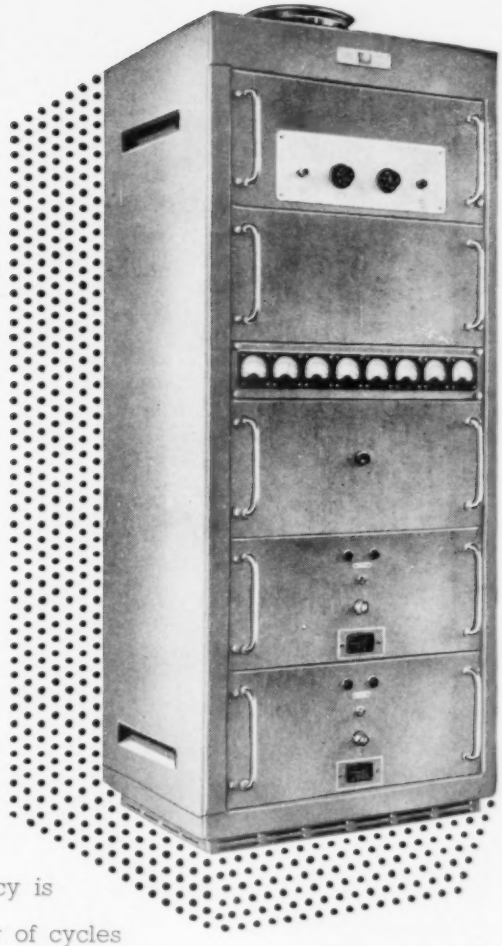

Designed for the measurement of any frequency in the range 10 c/s to 20 Mc/s with a basic accuracy of ± 1 part in 10^6 ± 0.1, 1.0, or 10 c/s. Higher accuracies available if required. The unknown frequency is determined by counting the number of cycles that pass through a 'gate' open for a selectable time interval of 0.1, 1.0, or 10 seconds. The result is presented on eight panel mounted meters each scaled 0 to 9 and is in decimal notation. Full information available on request.

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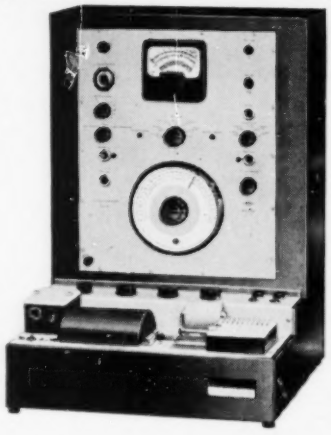
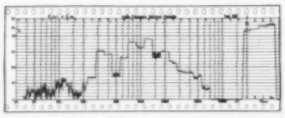


Analysis of an automobile horn



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The main tables give the electrical characteristics of each valve, and separate tables show their base connections. Sub-divisions of the main tables further classify the valves into obsolete, replacement or current types, as recommended by the makers. An index enables any valve to be found in the tables quickly and without trouble. A new feature is the comprehensive list of equivalents.

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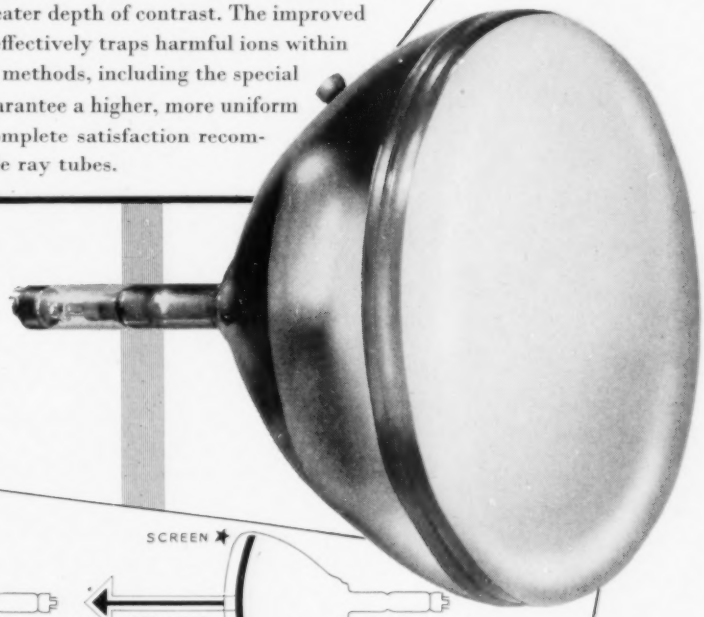

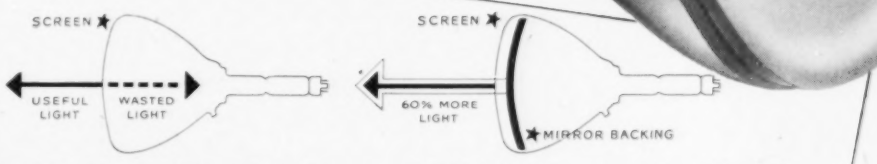

Without Aluminizing

Without Aluminizing, tubes waste half their light by backward reflection. To counteract this the brilliance must be increased which shortens tube life.

With Ediswan Aluminizing

Ediswan aluminized tubes have a mirror backing to the screen. All the light is thus thrown forward giving brighter, clearer pictures, and extra life.

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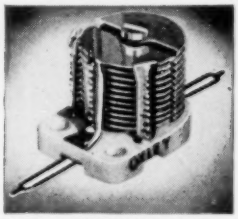

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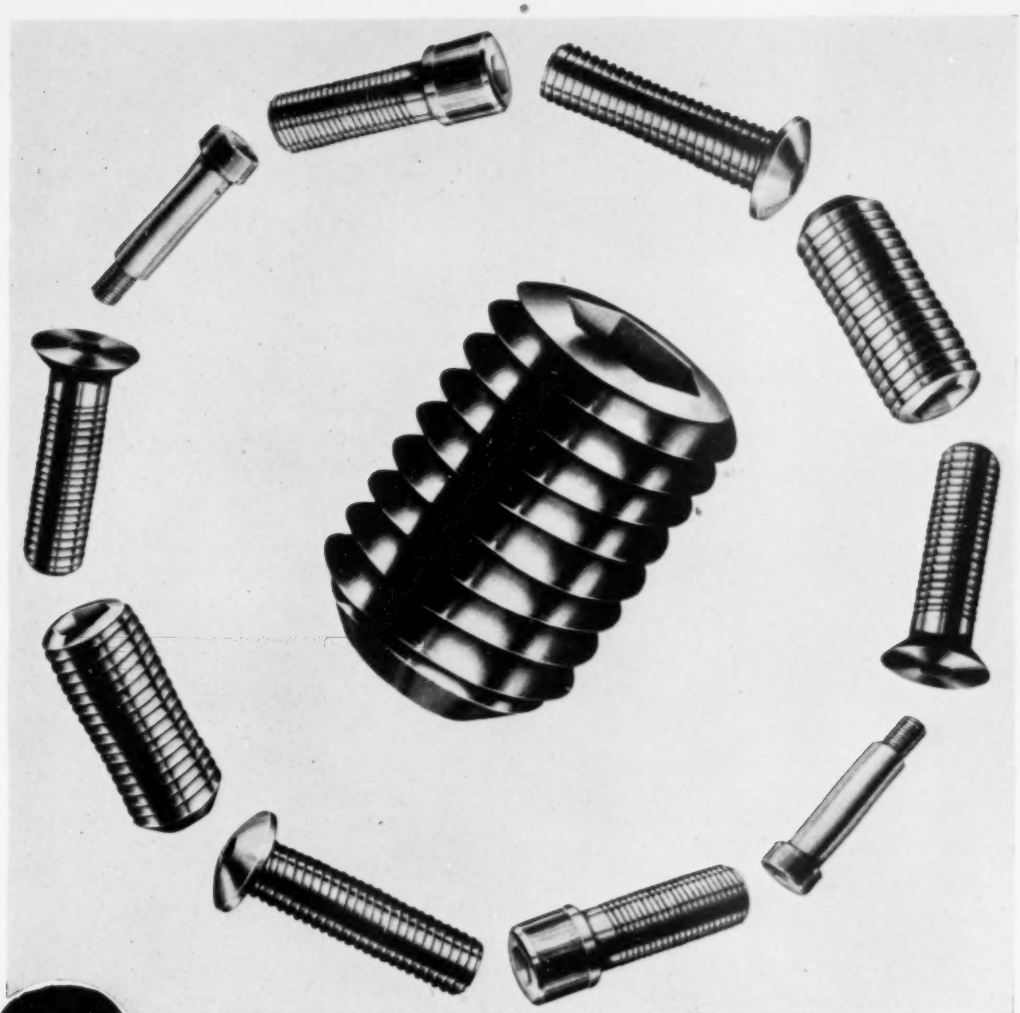

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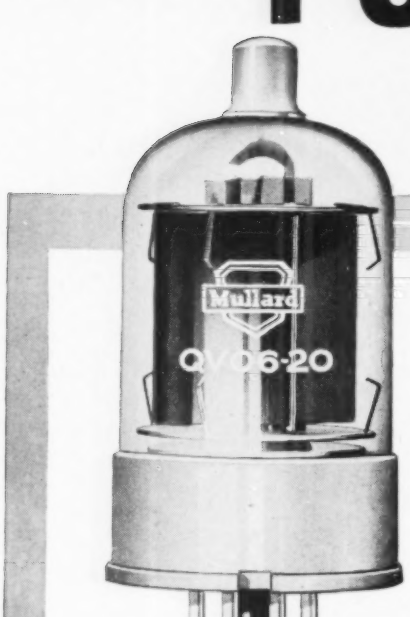

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WIRELESS ENGINEER

Vol. 31 AUGUST 1954 No. 8

Negative Ions in Cathode-Ray Beams

IT is nearly twenty years since von Ardenne, who had come up against this problem in the practical development of the cathode-ray tube, published a paper on the subject in *Arch. für Elektrotech.* (p. 731, 1935), and there have since been a number of papers dealing with various aspects of the subject. *Philips Research Reports* for April 1954 contains an article in German by W. F. Niklas in which the subject is discussed and experiments described which were made with tubes specially designed for the purpose. The beam leaving the cathode contains a number of negative ions in addition to the electrons, and if the focusing is purely electrostatic they will both be equally focused on the screen. If, in such a tube, the ions are not so sharply focused, they must have originated from another source such as gaseous collision, and not from the cathode. This does not apply to a magnetic-lens system, which has a negligible effect on the ions. If the deflecting system of a television tube is electromagnetic it has little effect on the ions, which consequently strike the screen at its centre and deactivate it, causing a dark spot called the ion burn. The article referred to discusses the origin of the ions and the ways in which they can be minimized or rendered harmless.

The experimental tube is shown in Fig. 1; it has an electrostatic focusing system producing a sharply-focused beam of negative ions; just where the beam emerges from the narrow tube, there is a magnet (7) producing a field of about 2 Wb/cm² which deflects the ions to an extent depending on their mass. (9) indicates the path of the electrons which are deflected by the stray field of the magnet (7) and thus eliminated from the beam. The screen material is a mixture of zinc sulphide and zinc-cadmium sulphide, between which and the glass is a transparent metallic layer deposited

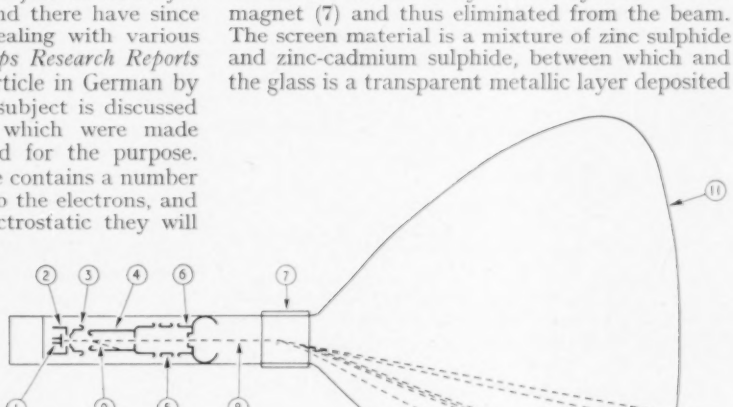

Fig. 1.

on the glass in order to prevent the setting up of any potential differences which might distort the spectrum. The voltage $V_3 = 300$ V and $V_4 = V_6 = 15{,}000$ V; V_5 is adjusted to focus the beam on the screen with a current of about 50 μA.

In order to interpret the spectra on the screen it is necessary to calibrate a tube. This was done

by dropping HCl on the Ba-Sr carbonate cathode and then in the closed tube converting the carbonate into oxide. A and B in Fig. 2 show the resulting spectra; it was decided that point 1 corresponds to Cl^{37} and point 2 to Cl^{35}; the greater intensity of point 2 agreed with the isotope ratio, viz. 70 to 80% Cl^{35} and 30 to 20% Cl^{37}. If m and m_x are the masses of two ions and d and d_x the distances of the spectrum points from the centre of the screen then $m_x/m = d^2/d_x^2$. Ions emanating from the cathode should all be focused with equal sharpness; a badly-focused point indicates that the relevant ions originated elsewhere. In Fig. 2 the points 1, 2 and 3 are sharp, whereas points 4, 5 and 6 are blurred (this is not very clear in the reproduction) and the author associates the former

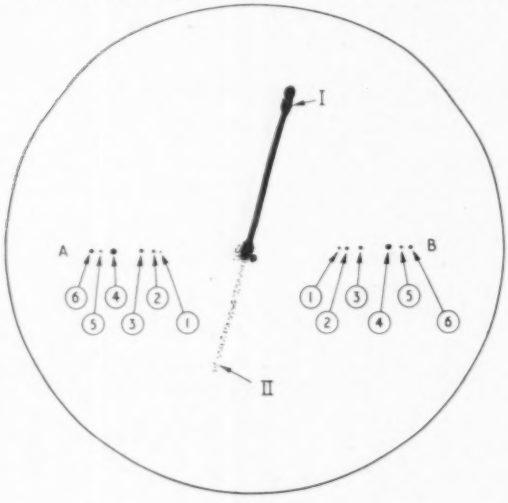

Fig. 2.

with cathode-ions and the latter with gas-ions. The pressure in the tube was about 5×10^{-7} cm. The gas-ions may have originated anywhere between the cathode and the accelerating anode, and therefore cannot focus at the same point, but most of them will be produced between the cathode and the low-voltage anode (3) where the electrons are moving slowly. A tube with specially arranged electrodes was made to test this point and it was concluded that most of the gas-ions were produced between the Wehnelt cylinder (2) and the 300-V anode (3). They would therefore attain a velocity only 2% less than those originating at the cathode.

TABLE

Point	Mass	Composition	Source
1	37	Cl^{37}	Cathode
2	35	Cl^{35}	Cathode
3	26	CN, C_2H_2	Cathode
4	16	O	Gas
5	13	CH	Gas
6	12	C	Gas

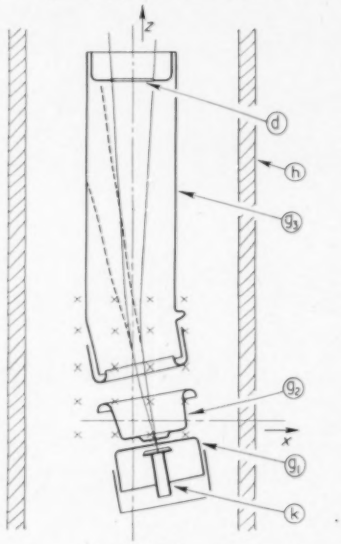

Fig. 3.

From the spectra shown in Fig. 2 the data corresponding to the six points are shown in the table.

If the vacuum is improved by means of a getter the points 4, 5 and 6 disappear when the pressure falls below 10^{-7} cm. If, however, gas is liberated by heating the metal parts of the gun until the pressure rises to 10^{-5} cm, the spectrum obtained is shown as I and II in Fig. 2. Part I is in the direction corresponding to negative ions, and Part II to positive ions. The continuity of the

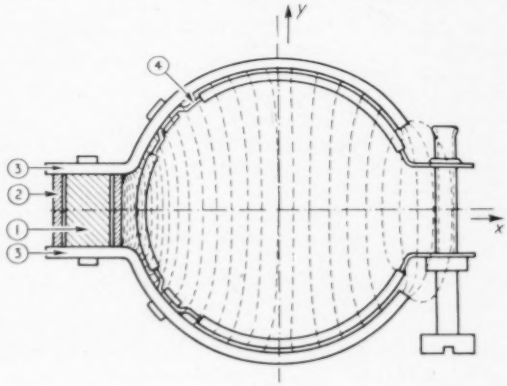

Fig. 4.

spectrum instead of separate points is explained by the continual ionization and the production of tertiary negative ions all along the path of the beam. The secondary positive ions which are responsible for the faint Part II of the spectrum are produced between electrodes (4) and (5) where the electric field is in the direction to accelerate them. Returning to the better vacuum, it is found that after the tube has been running for some hours, points 4, 5 and 6 and also point 3 become fainter; this the author ascribes to the continued action of the getter.

The gas-ions can thus be practically eliminated by thoroughly evacuating the tube or by means of an efficient getter; both gas- and cathode-ions can be reduced by preliminary running of the tube. In his 1935 paper, von Ardenne recommends deflecting the beam electrostatically out of the visible field during the formation period, thus protecting the screen. Niklas says, however, that it is not possible by simple means to remove the Cl components of the spectrum, since the chlorine may be produced by the glass of the tube. The author gives a large number of references to earlier work on the subject; one of these, entitled "negative-ion components in the cathode-ray beam", by Bachman and Carnahan, in *Proc. Inst. Radio Engrs* of 1938, covers much the same ground and provides a useful introduction to the paper by Niklas.

In another paper by the same author, entitled "An Improved Ion-trap Magnet" and published in *Philips Technical Review* for February-March 1954, a well-known device is described for removing the ions from the beam. In the experimental tube in Fig. 1 the electrons were deflected along the path (9) by the magnetic field and thus removed from the beam; in the ion-trap the axis of the electrodes is bent so that the deflected stream of electrons is utilized and the undeflected ions are trapped. This is shown in Fig. 3 where the ions follow the dotted path and strike the inner surface of the electrode while the deflected electrons pass through the opening and along the axis of the tube. The angle of deflection is 11° and it is important that the magnetic field distribution should be such that all the electrons are equally deflected so as to defocus the beam as little as possible. Fig. 4 shows the ordinary type of magnet employed for this purpose and Fig. 5 the new type. In both cases the Ticonal E permanent magnet (1) is surrounded by a bush (2) against which the soft-iron pole pieces (3) are clamped. The device is clamped on the tube by

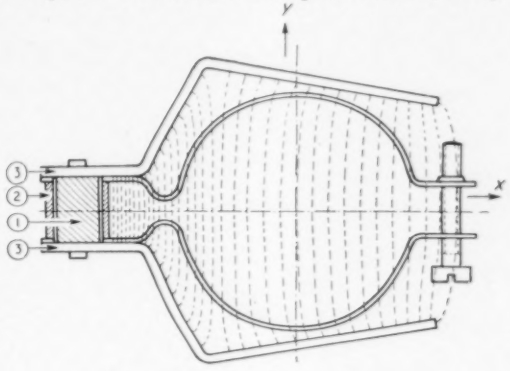

Fig. 5.

means of a non-magnetic padded strap (4) with a screw clamp so that its position can be adjusted. Since the angle of deflection depends on the strength of the magnetic field, it is essential that the strength of the field be uniform over the cross-section of the beam, and it was in order to obtain a more uniform field that the design of the magnet was changed to that shown in Fig. 5. The author says that the quality of the light spot is noticeably improved.

Acknowledgment is made to *Philips Research Reports* and to *Philips Technical Review* for permission to reproduce the illustrations.

G. W. O. H.

"WIRELESS ENGINEER" EDITORIALS

An index to the *Wireless Engineer* Editorials written by Professor G. W. O. Howe has been prepared by Dr. A. J. Small of Glasgow University. It covers the period January 1926 to May 1954 and is in three sections:—Index of titles, chronological; Index of subjects, and Index of names and authors.

As the titles of the Editorials do not fully describe all the subjects treated in them, the Subject Index is fully cross-referenced.

The index comprises 75 pages of quarto size with a limp paper binding, and is obtainable from Dr. A. J. Small, Electrical Engineering Department, The University, Glasgow, W.2, and costs 5s. including postage.

VAN DER POL'S EQUATION

Analytic Method of General Solution

By Ziya Akcasu

(Electronics Department, University of Southampton)

SUMMARY.—A new method for solving a quasi-linear equation of the form $x'' + x = \mu f(x)x'$ is proposed. The solution of the first approximation is obtained; it is compared with those given by others[1,2,3,4] and found to be identical with them. The method is applied then to the second approximation and yields the interdependence between the amplitude and instantaneous frequency during the transient period. Finally, the effect of the harmonic content is calculated in the transient state as well as in the steady-state. The results are applied to a thermionic generator.

1. Introduction

IN this paper an attempt is made to solve a non-linear equation of the form

$$\frac{d^2x}{dt^2} + x = \mu f(x)\frac{dx}{dt} \quad .. \quad .. \quad (1)$$

where μ is a real number assumed to be small compared with unity, and where $f(x)$ is an analytical function of x. This equation represents a typical and general class of non-linear differential equations encountered in electrical circuits containing a non-linear dissipative element. In many applications, however, the original equation obtained directly from the circuit is of the integral form of the above, viz.,

$$\frac{dx}{dt} + \int x\,dt = \mu F(x) \quad .. \quad .. \quad (2)$$

where

$$F(x) = \int f(x)dx \; .. \quad .. \quad .. \quad (3)$$

$F(x)$ represents the current–voltage characteristic of the non-linear resistive element. As will be seen from the application given at the end of this paper, no constant term appears in this original equation. But generally this integro-differential equation is not suitable because of the integral term. Therefore, it is usually differentiated with respect to t. This leads to the equation (1) which is a special type of the more general non-linear differential equation

$$\frac{d^2x}{dt^2} + x = \mu f\left(x, \frac{dx}{dt}\right) .. \quad .. \quad (4)$$

For this equation there exist several quantitative methods yielding approximate solutions.

In these methods it is generally assumed that the solution of the equation is of the form

$$x = a \sin (t + \theta) \quad .. \quad .. \quad (5a)$$

where a and θ are unknown functions of t. (For the sake of simplicity of expressions one sets $\omega = 1$, hence works with the normalized frequency; see later.) It is obvious that this is equivalent to

$$x = b_1 \sin t + b_2 \cos t \; .. \quad .. \quad (5b)$$

where b_1 and b_2 are again functions of t. Either of these solutions is substituted into the equation (1) and by this means two auxiliary equations are obtained from which a, θ or b_1, b_2 can be found.

In this paper the same procedure will be followed, but the solution of 5(a) or 5(b) will be substituted into the equation (2) instead of into equation (1).

MS accepted by the Editor, October 1953

2. First Approximation

If the solution of the form 5(a) is substituted into (2) the first term becomes

$$\frac{dx}{dt} = \frac{da}{dt}\sin(t + \theta) + a\left(1 + \frac{d\theta}{dt}\right)\cos(t + \theta) \quad (6)$$

For the second term the integral $\int a \sin (t + \theta)dt$ must be evaluated. For this case, however, it is more convenient to use the solution of the form 5(b); e.g.,

$$\int (b_1 \sin t + b_2 \cos t)dt = -b_1 \cos t + b'_1 \sin t - \int b''_1 \sin t\,dt + b_2 \sin t + b'_2 \cos t - \int b''_2 \cos t\,dt \quad .. \quad .. \quad .. \quad (7a)$$

or

$$\int [(b_1 + b''_1)\sin t + (b_2 + b''_2)\cos t]dt = (-b_1 + b'_2)\cos t + (b_2 + b'_1)\sin t \quad (7b)$$

where the dashes show the derivatives with respect to t. So far no assumption has been made about b_1 and b_2. Now it is assumed that b_1 and b_2 vary slowly with time. As will be seen later, the second derivatives of b_1 and b_2 are of the order of μ^2. As a first approximation b''_1 and b''_2 will be neglected in comparison with b_1 and b_2.

Substituting $b_1 = a\cos\theta$, $b_2 = a\sin\theta$ into (7) there results

$$\int a\sin(t + \theta)dt = \left(-a + a\frac{d\theta}{dt}\right)\cos(t + \theta) + \frac{da}{dt}\sin(t + \theta) \quad .. \quad .. \quad .. \quad (8)$$

The third term gives:

$$F[a \sin (t + \theta)] = \phi_1(a) \sin (t + \theta) + \phi_2(a) \sin [2(t + \theta)] + \ldots + \psi_0(a) + \psi_1(a) \cos (t + \theta) + \psi_2(a) \cos [2(t + \theta)] + \quad (9)$$

where ϕ_i, ψ_i are Fourier coefficients of the function $F(a \sin u)$.

By combining the equations (2), (6), (8) and (9) and equating the coefficients of $\sin (t + \theta)$ and $\cos (t + \theta)$ one obtains

$$\frac{da}{dt} = \frac{\mu}{2}\phi_1(a) \quad \ldots \quad (10)$$

$$\frac{d\theta}{dt} = \frac{\mu}{2}\psi_1(a) \quad \ldots \quad (11)$$

where

$$\phi_1(a) = \frac{1}{2\pi}\int_0^{2\pi} F(a \sin u) \sin u \, du \quad \ldots \quad (12a)$$

$$\psi_1(a) = \frac{1}{2\pi}\int_0^{2\pi} F(a \sin u) \cos u \, du = 0 \quad (12b)$$

Here all the harmonic terms have been disregarded.

The equations (10) and (11) are the auxiliary equations from which a and θ are to be solved. From (12b) it follows that $d\theta/dt = 0$; this means that the instantaneous frequency does not change with time.

The equations (10) and (11) are identical with those obtained in a different way by Kryloff and Bogoliuboff.[2] Nothing will be said in this paper about the stability of this solution since this question is widely investigated in the book just mentioned and also in that by Minorsky.[3] But it must be recalled that the equations are obtained by ignoring second derivatives in (7b) and the harmonics in (9). The constant term in (9) represents the well-known detection current due even power terms in $F(x)$. This is also ignored since only the oscillatory solution is considered in this treatment.

3. Second Approximation

To illustrate the effect of the second derivatives of the amplitude, the normal linear differential equation will be first discussed. If it is assumed, that in the equations (2) and (9) $F(x) \equiv x$, then from equations (10) and (11) there follows:

$$\frac{da}{dt} = \frac{\mu}{2}a; \quad \frac{d\theta}{dt} = 0 \quad \ldots \quad (13)$$

If a and θ are solved from these equations and substituted in (5a) one obtains

$$x = C\, e^{\mu t/2} \sin (t + \theta_0) \quad \ldots \quad (14)$$

where C and θ_0 are the constants of integration. On the other hand the exact solution of the linear differential equation

$$x'' + x = \mu x'$$

is known to be

$$x = Ce^{\mu t/2} \sin \left[\left(\sqrt{1 - \frac{\mu^2}{4}}\right)t + \theta_0\right] \quad (15)$$

The two solutions (15) and (16) differ in the frequency. This difference cannot be due to the neglect of harmonics, because there are no harmonics in a linear circuit. Hence the discrepancy can only be due to the neglect of the second derivative in equation (7b). To prove this point a solution of the form $x = a \sin (\omega t + \theta_0)$ is substituted in the above linear equation. In this solution ω and θ_0 are constants and a is a function of time which can easily be determined.

The result is

$$\frac{da}{dt} = \frac{\mu}{2}a$$

$$\omega = \sqrt{1 - \frac{1}{a}\frac{d^2a}{dt^2}} = \sqrt{1 - \frac{\mu^2}{4}} \quad \ldots \quad (16)$$

From this it follows that the frequency of linear oscillations depends on the second derivative of the amplitude. It also follows that the solution of the first approximation (14) is accurate as far as the amplitude is concerned.

If the second derivatives are to be taken into account, integration by parts must be carried one stage further than is done in (7a). Hence

$$\int (b_1 \sin t + b_2 \cos t)\, dt = -b_1 \cos t + b'_1 \sin t + b''_1 \cos t - \int b'''_1 \cos t \, dt + b_2 \sin t + b'_2 \cos t - b''_2 \sin t + \int b'''_2 \sin t \, dt \quad (17)$$

$$\int [(b_1 - b'''_2) \sin t + (b_2 + b'''_1) \cos t]\, dt = (-b_1 + b''_1 + b'_2) \cos t + (b'_1 + b_2 - b''_2) \sin t \quad \ldots \quad (18)$$

It is assumed that

$$\frac{b'''_2}{b_1} \ll 1 \quad \text{and} \quad \frac{b'''_1}{b_2} \ll 1$$

This means that the third derivatives can be neglected in comparison with the original functions. Substituting in (18) $b_1 = a \cos \theta$ and $b_2 = a \sin \theta$

$$\int a \sin (t + \theta) dt = (-a + a\theta' + a'' - a\theta'^2) \cos (t + \theta) + (a' - 2a'\theta' - a\theta'') \sin (t + \theta) \quad \ldots \quad (19)$$

The evaluation of this integral is given in Appendix I.

By combining the equations (2), (6), (9) and (19), and equating the coefficients of sin and cos terms one obtains

$$\frac{da}{dt} - \frac{1}{2}\frac{1}{a}\frac{d}{dt}\left(a^2\frac{d\theta}{dt}\right) = \frac{\mu}{2}\phi_1(a) \quad \ldots \quad (20)$$

$$\left(\frac{d\theta}{dt}\right)^2 - 2\frac{d\theta}{dt} = \frac{1}{a}\frac{d^2a}{dt^2} \quad .. \quad .. \quad (21)$$

Since the first approximation has been found sufficiently accurate for obtaining the amplitude, the second term in (20) will be ignored. The first term in (21) will also be neglected since $(d\theta/dt)^2$ is of the order of μ^4.

Instead of using equations (10) and (11) a and θ are solved from the equations

$$\frac{da}{dt} = \frac{\mu}{2}\phi_1(a) \; .. \quad .. \quad .. \quad .. \quad (22)$$

$$\frac{d\theta}{dt} = -\frac{1}{2}\frac{1}{a}\frac{d^2a}{dt^2} \quad .. \quad .. \quad .. \quad (23)$$

and the values obtained substituted into $x = a \sin(t + \theta)$. Thus an improved solution is found. The instantaneous frequency is obtained by differentiating $(t + \theta)$ with respect to t whence

$$\omega = 1 - \frac{1}{2}\frac{1}{a}\frac{d^2a}{dt^2} \quad .. \quad .. \quad .. \quad (24)$$

It must be realized that ω is the normalized frequency; i.e., $\omega_0^2 = 1/LC$ is supposed to be unity. By substituting d^2a/dt^2 from (22), one obtains

$$\omega = 1 - \frac{\mu^2}{8}\theta_1(a)\frac{d\phi_1(a)}{dt} \quad .. \quad .. \quad (25)$$

4. Effect of Harmonics

It is seen from the equation (25) that the influence of the amplitude variations on the frequency is of the order of μ^2. On the other hand, it has been shown by Kryloff and Bogoliuboff[2] that the reduction in the frequency due to the harmonics is also of the order of μ^2. It follows that to ignore the harmonics as done in the previous section is not justified if one is interested in frequency changes of the order of μ^2. It is the purpose of this section to take into account the harmonics as well as the second derivatives.

For the sake of simplicity it will be assumed that $F(x)$ is an odd function of x; i.e., that $F(x) = -F(-x)$. In this case the Fourier series of $F(a \sin u)$ contains, as can easily be seen, only odd sine terms. Hence

$$F(a \sin u) = \phi_1(a) \sin u + \phi_3(a) \sin 3u + \phi_5(a) \sin 5a + \ldots \quad .. \quad .. \quad (26)$$

It is also assumed that in the solution only the third harmonic is pronounced. Then the solution will be of the form

$$x = a \sin(t + \theta) + b \cos[3(t + \alpha)] \; .. \quad (27)$$

where a, θ, b and α are functions of time. It must be borne in mind that $d\alpha/dt$ may not be equal to $d\theta/dt$ during the transient period whereas it is known to be so for the stationary oscillations.

As done previously, the expression (27) for x must be substituted in equation (2).
Then the first term is

$$\frac{dx}{dt} = \frac{da}{dt}\sin(t + \theta) + a\left(1 + \frac{d\theta}{dt}\right)\cos(t + \theta) + \frac{db}{dt}\cos[3(t + \alpha)] - 3b\left(1 + \frac{d\alpha}{dt}\right)\sin[3(t + \alpha)] \quad .. \quad .. \quad .. \quad .. \quad (28)$$

With the help of the formula given in Appendix 1 the second term becomes

$$\int x dt = (-a + a\theta' + a'')\cos(t + \theta) + a \sin(t + \theta) + \frac{b}{3}(1 - \alpha')\sin[3(t + \alpha)] + \frac{b'}{9}\cos[3(t + \alpha)] \; .. \quad .. \quad .. \quad (29)$$

When evaluating this integral the second derivatives of the third harmonic are neglected because these quantities which are small in themselves appear with a factor 1/9.

The third term in equation (2) requires a comparatively longer calculation. At this stage it is assumed that the amplitude of the third harmonic is small in comparison with that of the fundamental. It follows from this that

$$F[a \sin(t + \theta) + b \cos 3(t + \alpha)] \simeq F[a \sin(t + \theta)] + \frac{dF[a \sin(t + \theta)]}{d[a \sin(t + \theta)]} b \cos 3[(t + \alpha)] \quad .. \quad .. \quad .. \quad .. \quad (30)$$

If the function $\dfrac{dF(a \sin u)}{d(a \sin u)}$ is expanded into a Fourier series

$$\frac{dF(a \sin u)}{d(a \sin u)} = \psi_0(a) + \psi_2(a) \cos 2u + \psi_4(a) \cos 4u + \cdots \quad .. \quad (31)$$

The coefficients can be calculated in terms of those that are already determined in equation (26). If the equation (26) is differentiated with respect to u one obtains

$$\frac{dF(a \sin u)}{du} = \frac{d[F(a \sin u)]}{d(a \sin)\, u} a \cos u = \phi_1(a) \cos u + 3\phi_3(a) \cos 3u + 5\phi_5(a) \cos 5u + \cdots$$

Dividing this by $(a \cos u)$, evaluating the terms

$$\frac{\cos 3u}{\cos u}, \quad \frac{\cos 5u}{\cos u} \ldots,$$

and combining it with (31) gives (see Appendix 2)

$$\left.\begin{aligned} \psi_0(a) &= \frac{1}{a}[\phi_1(a) - 3\phi_3(a) + 5\phi_5(a) \; \ldots] \\ \psi_2(a) &= \frac{2}{a}[3\phi_3(a) - 5\phi_5(a) + \ldots] \\ \psi_4(a) &= \frac{2}{a}[5\phi_5(a) - 7\phi_3(a) + \ldots] \end{aligned}\right\} \quad (32)$$

Combining the equations (27), (31) and (2) and denoting $3(\theta - \alpha)$ by δ gives

$$F(x) = \left[\phi_1(a) + \frac{b}{2}[\psi_2(a) - \psi_4(a)] \sin\delta\right] \sin(t+\theta) + \left[3\frac{b}{a}\phi_3(a)\cos\delta\right]\cos(t+\theta) + \left[b\psi_0(a) + \frac{b}{2}\psi_6(a)\cos 2\delta + \phi_3(a)\sin\delta\right]\cos 3(t+\alpha) + \left[\phi_3(a)\cos\delta - \frac{b\psi_6(a)}{2}\sin 2\delta\right]\sin 3(t+\alpha) \quad .. \quad .. \quad .. \quad (33)$$

Finally, the equations (28), (29) and (33) are substituted into (2) and sin and cos terms are equated. Then

$$\frac{d\theta}{dt} = -\frac{1}{2a}\frac{d^2a}{dt^2} + \frac{3}{2}\mu\frac{\phi_3(a)}{a}\cos\delta \quad .. \quad .. \quad (34)$$

$$\frac{da}{dt} = \frac{\mu}{2}\left[\phi_1(a) + \frac{b}{a}[\psi_2(a) - \psi_4(a)]\sin\delta\right] \quad .. \quad (35)$$

$$b\left(\frac{4}{5} + \frac{d\alpha}{dt}\right) = -\frac{3}{10}\mu\phi_3(a)\cos\delta + \frac{\mu}{2}b\psi_6(a)\sin 2\delta \quad .. \quad .. \quad .. \quad .. \quad (36)$$

$$\frac{db}{dt} = \frac{9}{10}\mu\left[b\psi_0(a) + \phi_3(a)\sin\delta - \frac{b}{2}\psi_6(a)\cos 2\delta\right] \quad .. \quad .. \quad .. \quad .. \quad (37)$$

From these equations a, θ, b and α can be found. In equation (35) the last term shows the effect of the third harmonic on the amplitude of fundamental frequency. As will be seen from the formulae (39) and (42), obtained below for b and δ, the magnitude of the second term is of the order of μ^2 and, therefore, can be neglected.

The same applies to $d\alpha/dt$ and to the last term in (36). Hence equation (35) gives

$$\frac{da}{dt} = \frac{\mu}{2}\phi_1(a) \quad .. \quad .. \quad .. \quad (38)$$

from which a can be found as a function of time, and (36) gives b in terms of a; i.e.,

$$b = -\tfrac{3}{8}\mu\phi_3(a) \quad .. \quad .. \quad (39)$$

Here $\cos\delta$ is set equal to unity. The error due to this approximation is of the order of μ^2 because δ is proportional to μ.

The change in the frequency of the fundamental is obtained by combining the equations (34) and (39), whence

$$\frac{d\theta}{dt} = -\frac{1}{2a}\frac{d^2a}{dt^2} - 4\frac{b^2}{a^2} \quad .. \quad (40)$$

The first term on the right-hand side of (40) is the change due to the varying amplitude and therefore applies only to the transient period. The second term shows the effect of the third harmonic.

The change in the frequency of the third harmonic (i.e., $d\alpha/dt$) will now be found.

It seems appropriate to evaluate $d\delta/dt$ rather than $d\alpha/dt$, because the former gives, during the transient period, the deviation of the 'third harmonic' from three times the fundamental frequency. For the stationary oscillations it is known that this deviation is zero. It must be recalled that δ was equal to $3(\theta - \alpha)$ and consequently

$$\frac{d\delta}{dt} = 3\left(\frac{d\theta}{dt} - \frac{d\alpha}{dt}\right) \quad .. \quad .. \quad (41a)$$

or

$$\frac{d\delta}{dt} = 3\omega_f - \omega_h \quad .. \quad .. \quad (41b)$$

where the subscripts f and h refer to fundamental and harmonic respectively.

By differentiating equation (39) with respect to t and substituting da/dt from (38) one obtains

$$\frac{db}{dt} = -\frac{3}{16}\mu^2\frac{d\phi_3(a)}{da}\phi_1(a)$$

Combining this with (37) gives

$$\delta = \tfrac{3}{8}\mu[\phi_1(a) - 3\phi_3(a) + 5\phi_5(a)] - \frac{5}{24}\mu\frac{\phi_1(a)}{\phi_3(a)}\frac{d\phi_3(a)}{da} \quad .. \quad .. \quad (42)$$

With the help of the equations (38), (39), (40) and (42) one can determine the unknown quantities in (27) and thus obtain the solution required.

5. Application

Consider a valve oscillator with a tuned anode circuit (Fig. 1). The currents in the

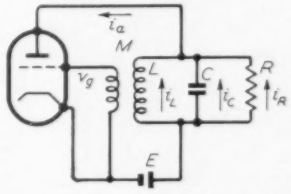

Fig. 1. *Valve oscillator with tuned-anode circuit.*

circuit are expressed in terms of the p.d. v across C as follows:—

$$i_R = \frac{v}{R};\ i_C = C\frac{dv}{dt};\ I_L = I_0 + i_L = I_0 + \frac{1}{L}\int v.dt \quad I_a = \psi(v_a + \mu v_g) \quad (43)$$

where I_0 is the d.c. component of the current in the inductance and where the function $\psi(v_a + \mu v_g)$ shows the non-linear valve characteristic. v_a stands for the effective anode voltage; i.e., $v_a = E - v$.

Remembering that $v_g = \frac{M}{L}v$ one can write I_a as follows

$$I_a = \psi(E + kv)$$

where $\kappa = \left(\mu \frac{M}{L} - 1\right)$

Kirchhoff's equation for the currents gives

$$i_c + i_R + I_L = I_a$$

or

$$i_c + i_R + i_L = I_a - I_0 = i_a \qquad (44)$$

where i_a and i_L are the varying components of the currents in the valve and in the inductor respectively.

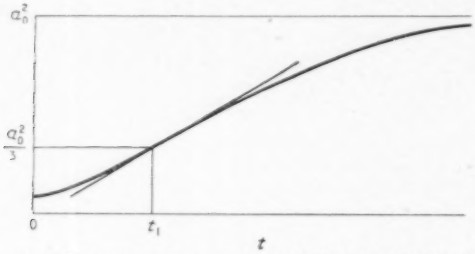

Fig. 2. *This curve represents equation (51) for the triode oscillator under consideration.*

Since E is a constant voltage, i_a can be expressed as a function of v; i.e.,

$$i_a = \phi(v) \qquad .. \quad .. \quad .. \qquad (45)$$

By substituting from (43) into (44) and denoting

$$\frac{1}{\omega_0 C}\left[\phi(v) - \frac{v}{R}\right] \text{ by } \mu F(v) \text{ one obtains}$$

$$\frac{dv}{dt} + \int v dt = \mu F(v) \qquad .. \quad .. \quad .. \qquad (46)$$

where t is the normalized time; i.e., $t = \omega_0 \tau$, and $\omega_0^2 = 1/LC$.

It is assumed now that the characteristic of the non-linear resistive element takes the form

$$F(v) = v(1 - \gamma v^2) \qquad .. \quad .. \qquad (47)$$

For $v = a \sin u$ (47) becomes

$$F(a \sin u) = a(1 - \tfrac{3}{4}\gamma a^2) \sin u + \frac{\gamma a^3}{4} \sin 3u \qquad (48)$$

Thus, the coefficients in equation (26) are now

$$\left.\begin{aligned} \phi_1(a) &= a\left(1 - \frac{a^2}{a_0^2}\right) \\ \phi_3(a) &= \frac{1}{3}\frac{a^3}{a_0^2} \end{aligned}\right\} \qquad .. \quad .. \qquad (49)$$

where $a_0^2 = 4/3\gamma$.

From (38) it follows that

$$\frac{da}{dt} = \frac{\mu}{2} a\left(1 - \frac{a^2}{a_0^2}\right) \qquad .. \quad .. \qquad (50)$$

From (50)

$$a = \frac{a_0}{\sqrt{1 + c a_0^2 e^{-\mu t}}} \qquad .. \quad .. \qquad (51)$$

where c is the constant of integration. For large values of t, a approaches a_0, the amplitude of the steady-state oscillation.

For the amplitude of the third harmonic, equation (39) gives

$$b = -\frac{\mu}{8}\frac{a^3}{a_0^2} \qquad .. \quad .. \quad .. \qquad (52)$$

The instantaneous angular frequency of the fundamental is obtained from (40) as follows

$$\omega = 1 + \frac{d\theta}{dt} = 1 - \frac{1}{2a}\frac{d^2a}{dt^2} - \frac{\mu^2}{16}\frac{a^4}{a_0^4} \qquad .. \qquad (53)$$

or, evaluating d^2a/dt^2 from (50),

$$\omega = 1 - \frac{\mu^2}{8}\left(1 - \frac{a^2}{a_0^2}\right)\left(1 - 3\frac{a^2}{a_0^2}\right) - \frac{\mu^2}{16}\frac{a^4}{a_0^4} \qquad .. \quad .. \quad .. \qquad (54)$$

It must be recalled that ω shows the normalized frequency. The curves representing equations (51) and (54) are shown in Figs. 2 and 3 respectively. From the graph it is seen that the frequency of oscillation is at first smaller and then larger than the steady-state value.

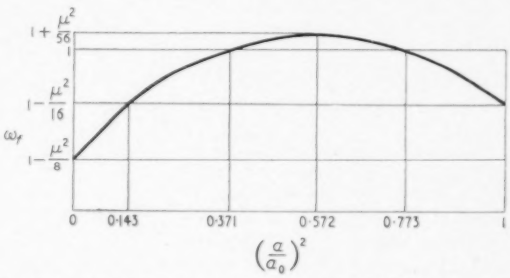

Fig. 3. *Equation (54) produces this curve for the example considered.*

To find the frequency of the '3rd harmonic', equation (42) must be investigated. By substituting the terms $\phi_1(a)$ and $\phi_3(a)$ from (49) into (42) one obtains

$$\delta = -\frac{\mu}{8}\left(2 + \frac{a^2}{a_0^2}\right) \qquad .. \quad .. \qquad (55)$$

Differentiating (55) with respect to t,

$$\frac{d\delta}{dt} = 3\omega_f - \omega_h, \text{ viz.,}$$

$$\frac{d\delta}{dt} = -\frac{\mu^2}{8}\frac{a^2}{a_0^2}\left(1 - \frac{a^2}{a_0^2}\right) \qquad .. \qquad (56)$$

Fig. 4 shows $-\frac{d\delta}{dt}$ as a function of $\left(\frac{a^2}{a_0^2}\right)$. It follows that the frequency of the '3rd harmonic' is slightly larger than three times that of the fundamental during transient time.

Fig. 5 represents the amplitude of the '3rd

harmonic' as a function of $(a/a_0)^2$ during transient time.

The solution for the steady-state oscillation is obtained by replacing a by a_0 in equations (52), (54) and (55).

Hence

$$x = a_0 \sin\left[\left(1 - \frac{\mu^2}{16}\right)t + \theta_0\right] - \frac{\mu a_0}{8}\cos 3\left[\left(1 - \frac{\mu^2}{16}\right)t + \theta_0 + \frac{\mu}{8}\right] \quad (57)$$

With the exception of the last term $\mu/8$ this solution is identical with that obtained by Kryloff and Bogoliuboff.

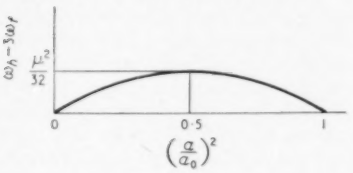

Fig. 4

Fig. 5

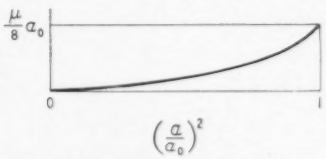

Fig. 4 (above). *Relation between $d\delta/dt$ and a^2/a_0^2.*

Fig. 5 (below). *Amplitude of 'third harmonic'.*

6. Conclusion

The use of the integro-differential equation has made it possible to deal more rigorously than before with conditions during transient period. The resultant formulae apply to the circuits containing the non-linear resistive element, whose current-voltage characteristic is an odd function. Amplitude and frequency of the fundamental and '3rd harmonic' are obtained both for the transient period and for the steady-state. The method can be applied to a resistive characteristic of a more general nature.

Acknowledgments

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APPENDIX 1

To evaluate the integral $\int a \sin(t + \theta)dt$ it is convenient to define a complex function $X = ae^{j\theta}$

Hence

$$\int a \sin(t + \theta)\,dt = \text{Imag.}\left[\int Xe^{jt}\,dt\right]$$

Integrating the latter by parts

$$\int Xe^{jt}\,dt = (-jX + X' + jX'')e^{jt} + j\int X'''e^{jt}\,dt.$$

Neglecting jX''' in comparison with X

$$\int Xe^{jt}\,dt = (-jX + X' + jX'')e^{jt}.$$

Substituting $X = ae^{j\theta}$ in the last equation

$$\int Xe^{jt}\,dt = [-ja + a' + ja\theta' + j(a'' + j2a'\theta' + ja\theta'' - a\theta'^2)]e^{j(t+\theta)}$$
$$= j[-a + a\theta' + a'' - a\theta'^2]e^{j(t+\theta)} + [a' - 2a'\theta' - a\theta'']e^{j(t+\theta)}$$

Retaining only imaginary terms gives

$$\int a \sin(t + \theta)\,dt = [-a + a\theta' + a'' - a\theta'^2]\cos(t + \theta) + [a' - 2a'\theta' - a\theta'']\sin(t + \theta)$$

APPENDIX 2

Let the function $F(a \sin u)$ be an odd function of $a \sin u$. Then

$$F(a \sin u) = \phi_1(a)\sin u + \phi_3(a)\sin 3u + \phi_5(a)\sin 5u + \ldots$$

Differentiation with respect to u gives

$$\frac{dF(a \sin u)}{du} = \frac{dF(a \sin u)}{d(a \sin u)}\,a \cos u = \phi_1(a)\cos u + 3\phi_3(a)\cos 3u + 5\phi_5(a)\cos 5u + \ldots$$

Dividing by $(a \cos u)$ gives

$$\frac{dF(a \sin u)}{d(a \sin u)} = \frac{\phi_1(a)}{a} + 3\frac{\phi_3(a)}{a}\frac{\cos 3u}{\cos u} + 5\frac{\phi_5(a)}{a}\frac{\cos 5u}{\cos u} + \ldots$$

On the other hand

$$\cos(2k + 1)u = \cos u \cos 2ku - \sin u \sin 2ku = \cos u \cos 2ku - \tfrac{1}{2}[\cos(2k - 1)u - \cos(2k + 1)u]$$

Hence

$$\frac{\cos(2k + 1)u}{\cos u} = 2\cos 2ku - \frac{\cos(2k - 1)u}{\cos u}$$

This recurrence formula gives for $k = 0, 1, 2 \ldots$

$$1 = 1$$

$$\frac{\cos 3u}{\cos u} = 2\cos 2u - 1$$

$$\frac{\cos 5u}{\cos u} = 2\cos 4u - 2\cos 2u + 1$$

$$\frac{\cos 7u}{\cos u} = 2\cos 6u - 2\cos 4u + 2\cos 2u - 1$$

Substituting these expressions in (1) one obtains

$$\frac{dF(a \sin u)}{d(a \sin u)} = \frac{1}{a}[\phi_1(a) - 3\phi_3(a) + 5\phi_5(a) - \ldots] + \frac{2}{a}[3\phi_3(a) - 5\phi_5(a) + 7\phi_7(a) - \ldots]\cos 2u + \frac{2}{a}[5\phi_5(a) - 7\phi_7(a) + 9\phi_9(a) \ldots]\cos 4u + \ldots$$

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IMPEDANCE TRANSFORMATION BY FOUR-TERMINAL NETWORKS

Use of Graphical Methods

By S. Fedida, B.Sc.(Hons.), A.C.G.I., A.M.I.E.E.

SUMMARY.—The purpose of this paper is to present a study of the impedance transformation occurring between the input and output terminals of a network, to discuss the corresponding geometrical relationships and to apply these relationships to the field of impedance measurement.

LIST OF SYMBOLS

$Z_{1s}(Z_{2s})$	Input impedance of four-terminal network at terminals 1(2), when terminals 2(1) are short-circuited.
$Z_{1\infty}(Z_{2\infty})$	Input impedance of four-terminal network at terminals 1(2), when terminals 2(1) are open-circuited.
$Z_s(Z_\infty)$	Short-circuit (open-circuit) input impedance of symmetrical networks.
$Z_1(Z_2)$	Image impedance at terminals 1(2) of network.
Z_0	Characteristic impedance of network.
$Z_{1L}(Z_{2L})$	Impedance of load connected to terminals 1(2) of network.
$Z_{1in}(Z_{2in})$	Input impedance at terminals 1(2) of network for an arbitrary impedance $Z_{2L}(Z_{1L})$.
K	Reflection coefficient of normalized impedance Z (see Section 1), defined as: $K = \frac{Z-1}{Z+1}$. When K is followed by a subscript it becomes the reflection coefficient of the impedance with the same subscript, unless otherwise specified in the text.

1. Introduction

A LINEAR four-terminal network can, in general, be regarded as an impedance transformer. Short lengths of transmission lines are commonly used as such and it is useful, when dealing with problems involving the use of these elements, to be clear about some of the possible geometrical interpretations of such transformations, particularly when impedance charts have to be used. Often these geometrical properties indicate methods by which impedance measurements can either be speeded up, made more accurate or self-checking.

Impedances will be represented geometrically according to the Argand diagram convention,[1] where the real part of an impedance is plotted along a horizontal axis and the imaginary part along a vertical axis.

In Fig. 1, for example, vector OP is made to represent an impedance Z, as indicated in the bracket to the right of point P. If Z should vary in magnitude, direction or both, point P will trace a curve, which will be denoted the locus of Z.

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All impedances will be assumed to be normalized in terms of some impedance Z', which is generally the purely real characteristic impedance of a transmission line, used for high-frequency measurements or as a transmission component. In some cases, however, Z' may be the internal impedance of a generator connected to the network as, for example, in Section 3.2. In all insertion-loss determinations both generator and line impedances are assumed to be equal and purely real.

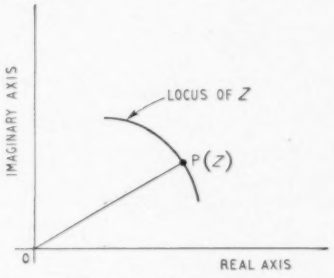


Fig. 1. *Illustration of the Argand diagram to represent impedances.*

The reflection coefficient of an impedance Z with respect to the normalizing impedance Z' is given by the well-known formula:

$$K = \frac{Z - Z'}{Z + Z'}$$

If the same symbol Z is now made to represent the same impedance normalized in terms of Z', then the above formula simplifies to:

$$K = \frac{Z - 1}{Z + 1} \qquad (1)$$

The normalizing impedance will always be chosen in such a way that equation (1) remains valid.

The reflection coefficient K will be treated as a vectorial quantity and it will be plotted as if it were an impedance, with the difference that its magnitude cannot exceed unity and its real part can take negative values. This type of plot is the well-known "Smith's Chart" diagram for transmission lines.

In order to avoid confusion it is convenient, in general, to plot the variations of load and input impedances on separate diagrams, even though these diagrams make use of the same co-ordinate systems.

We shall in this study make great use of the following set of formulae relating the input and load impedances of a four-terminal network, in terms of the short-circuit and open-circuit parameters of the latter:

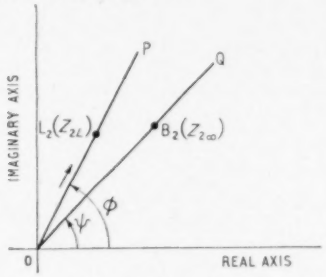


Fig. 2. *Load-impedance diagram; constant-phase locus.*

$$Z_{1in} = Z_{1\infty} + \frac{Z_{1\infty}(Z_{2s} - Z_{2\infty})}{Z_{2\infty} + Z_{2L}} \quad .. \quad (2)$$

$$Z_{2in} = Z_{2\infty} + \frac{Z_{2\infty}(Z_{1s} - Z_{1\infty})}{Z_{1\infty} + Z_{1L}} \quad .. \quad (3)$$

When the network is symmetrical (2) and (3) become

$$Z_{in} = Z_{\infty} + \frac{Z_{\infty}(Z_S - Z_{\infty})}{Z_{\infty} + Z_L} \quad .. \quad .. \quad (4)$$

Equations (2), (3) and (4) define a bilinear transformation[2,3] between the variables Z_{in} and Z_L and it can be proved that if one variable, say Z_L, varies in such a way that its locus is a circle (a straight line is considered to be a circle of infinite radius), then the locus of the other is also a circle.

This general theorem will be used throughout except when a direct derivation of the required result can be easily made.

2. Impedance Transformations in Cartesian Co-ordinates

We shall in this section analyse the types of input impedance loci obtained when the load impedance varies in certain specified ways.

The analysis is made simpler if equations (2), (3) and (4) are put in the following forms:

$$\frac{Z_{1in} - Z_{1s}}{Z_{1in} - Z_{1\infty}} = -\frac{Z_{2L}}{Z_{2\infty}} \quad .. \quad .. \quad .. \quad (5)$$

$$\frac{Z_{2in} - Z_{2s}}{Z_{2in} - Z_{2\infty}} = -\frac{Z_{1L}}{Z_{1\infty}} \quad .. \quad .. \quad .. \quad (6)$$

$$\frac{Z_{in} - Z_S}{Z_{in} - Z_{\infty}} = -\frac{Z_L}{Z_{\infty}} \quad .. \quad .. \quad .. \quad (7)$$

2.1. *Constant-Phase Circles*

Let us assume that the load impedance Z_{2L} varies in magnitude while keeping a constant phase ϕ, say, and let ψ be the phase angle of $Z_{2\infty}$.

We can write

$$Z_{2L} = |Z_{2L}|e^{j\phi} \text{ and } Z_{2\infty} = |Z_{2\infty}|e^{j\psi}$$

When these values are inserted in equation (5) we obtain

$$\frac{Z_{1s} - Z_{1in}}{Z_{1\infty} - Z_{1in}} = \frac{|Z_{2L}|}{|Z_{2\infty}|}e^{j(\pi+\phi-\psi)} \quad .. \quad (8)$$

Let L_2 and B_2 (Fig. 2) be the points in the load-impedance diagram representing Z_{2L} and $Z_{2\infty}$, and OP and OQ the radial lines through L_2 and B_2. OP represents a constant-phase locus for the impedance Z_{2L}.

If A_1, B_1 and C_1 (Fig. 3) represent the impedances Z_{1s}, $Z_{1\infty}$ and Z_{1in}, respectively, in the input-impedance diagram, then the vectors C_1A_1 and C_1B_1 represent the impedances $(Z_{1s} - Z_{1in})$ and $(Z_{1\infty} - Z_{1in})$, respectively. These two impedances can be expressed as follows

$$Z_{1s} - Z_{1in} = Ae^{j\alpha_s}, \quad Z_{1\infty} - Z_{1in} = Be^{j\alpha_\infty}$$

where A and B are positive real numbers and α_s and α_∞ are the angles shown in the figure.

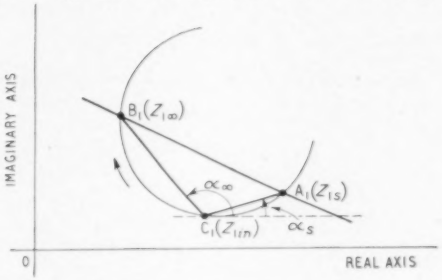


Fig. 3. *Input-impedance diagram: constant-phase locus.*

When these values are inserted in equation (8) we obtain

$$\frac{A}{B}e^{j(\alpha_s - \alpha_\infty)} = \frac{|Z_{2L}|}{|Z_{2\infty}|}e^{j(\pi+\phi-\psi)} \quad .. \quad .. \quad (9)$$

from which we deduce

$$\alpha_s - \alpha_\infty = \pi + \phi - \psi \quad .. \quad .. \quad .. \quad (10)$$

and consequently

$$\angle C_1B_1 . C_1A_1 = \pi + \angle OQ . OP \quad .. \quad (11)$$

where $\angle C_1B_1 . C_1A_1$ represents the angle through which it is necessary to rotate C_1B_1, around C_1, to make it coincide with C_1A_1. Similarly $\angle OQ . OP$ represents the angle through which it is necessary to rotate OQ to make it coincide with OP.

If now we increase the magnitude of Z_{2L}, starting from zero, while leaving its phase angle ϕ fixed, L_2 moves outward along OP, the angle

$\angle C_1B_1 . C_1A_1$, according to equation (11), remains constant and it follows that C_1 moves from A_1 to B_1 along an arc of circle.

Hence for every line of constant phase, such as OP, there corresponds in the input-impedance diagram an arc of circle, such as $A_1C_1B_1$, the locus of the transformed impedance.

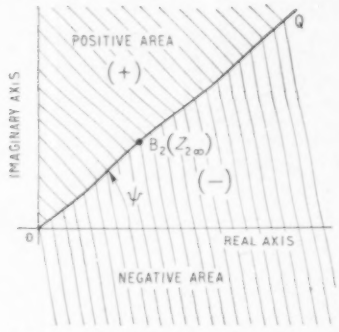


Fig. 4. *Load-impedance diagram; positive and negative areas.*

It is convenient at this point to divide the load-impedance diagram into two areas, with a common boundary at OQ (Fig. 4). The top area is such that $\phi - \psi$ is positive and it will consequently be called the positive area, while the bottom one will be called, for similar reasons, the negative area.

For every point in the positive area $\alpha_s - \alpha_\infty$ is, according to equation (10), greater than π radians and consequently the corresponding point in the input-impedance diagram lies below the line A_1B_1. Similarly, to points in the negative area correspond points above the line A_1B_1.

We may, therefore, divide the input-impedance diagram (Fig. 5) into two areas, with a common boundary at A_1B_1, corresponding to the two areas in the load impedance diagram.

For purely capacitive load impedances $\left(\phi = -\frac{\pi}{2}\right)$, $\alpha_s - \alpha_\infty$ is equal to $\frac{\pi}{2} - \psi$ and,

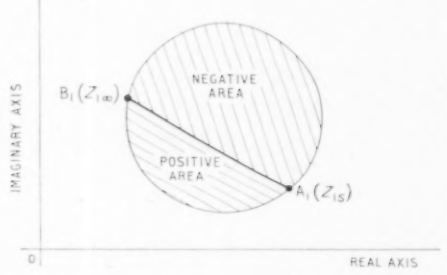


Fig. 5. *Input-impedance diagram; positive and negative areas.*

provided $\psi > 0$, it is smaller than $\pi/2$ radians; it follows that the centre of the arc of circle in the input-impedance diagram, corresponding to the negative imaginary axis, is in the negative area.

For purely inductive load impedances $\left(\phi = +\frac{\pi}{2}\right)$, $\alpha_s - \alpha_\infty = \frac{3\pi}{2} - \psi$ and consequently the arc of circle corresponding to the positive imaginary axis is a continuation of the arc of circle corresponding to the negative imaginary axis. The complete circle corresponding to the imaginary axis we shall call the reactance circle.

When the phase angle of the load impedance is between $-\frac{\pi}{2}$ and $+\frac{\pi}{2}$, the corresponding arcs of circle are contained within the reactance circle. As ϕ increases from $-\frac{\pi}{2}$ to $+\frac{\pi}{2}$, $\alpha_s - \alpha_\infty$ increases from $\frac{\pi}{2} - \psi$ to $\frac{3\pi}{2} - \psi$ and the point representing the input impedance transfers from the negative area to the positive area. When ϕ is equal to ψ, the constant-phase arc of circle coincides with the straight line A_1B_1.

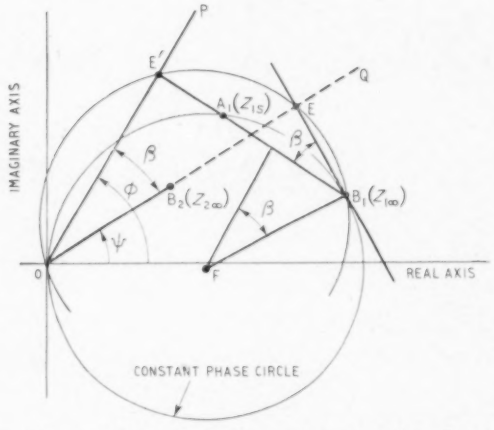


Fig. 6. *Input-impedance diagram; construction of constant-phase circle corresponding to line OP.*

Hence a four-terminal network transforms the right-hand side of the load-impedance diagram into the interior of the reactance circle, and the left-hand side of the load-impedance diagram into the remainder of the input-impedance diagram, outside the reactance circle.

The phase circle corresponding to a given load impedance can be found as follows: let OP (Fig. 6) be the phase line of the given load impedance, plotted in the input-impedance diagram, and circle of centre F the corresponding

phase circle. If B_1E is the tangent at B_1 to circle F, E its intersection with OQ and E′ the intersection of B_1A_1 with OP, then, since, according to equations (10) and (11), the angles shown as β in the figure are equal, it follows that points E, E′, E and B_1 are on a common circle and consequently point E can be determined by the intersection of OQ with the circle passing through O, E′ and B_1. The required phase circle passes through A_1 and B_1 and it is tangential at B_1 to EB_1. An alternative construction, shown in Fig. 7, makes use of the tangent at A_1 to the required phase circle, and of the intersection H of OQ and A_1B_1.

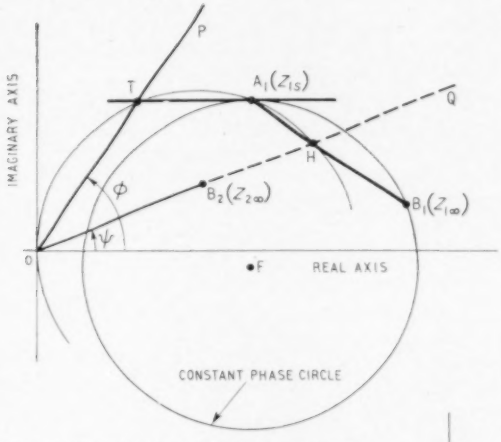


Fig. 7 (above). *Input-impedance diagram; alternative construction of constant-phase circle corresponding to line OP.*

Fig. 8 (right). *Input-impedance diagram; construction of constant-phase circle corresponding to line OP, for symmetrical networks.*

A much simpler construction can be used when the four-terminal network is known to be symmetrical (Fig. 8). If C_1 is any point on the required phase circle and C_2 the intersection of AC_1 and OP, then points O, C_1, B and C_2 are on the same circle, because of the equality of the angles shown as β in the figure.

Hence to find any point C_1 on the required phase circle, draw a circle, such as S, through B and O, intersecting the given phase line OP at C_2. C_1 is found at the intersection of AC_2 with circle S.

2.2. *Constant-Modulus Circles*

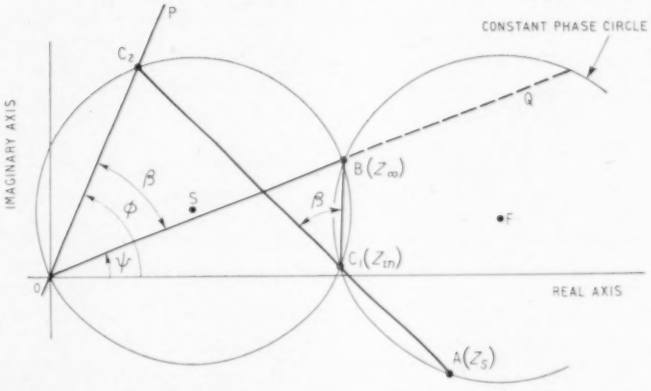
We have examined in the above paragraph the correspondence between input and load impedances of a four-terminal network, when the load impedance varies in such a way as to keep a constant-phase angle. We shall in this section examine the case of a load impedance, Z_{2L}, the magnitude of which remains constant.

The load-impedance diagram is obviously a semi-circle centred at the origin.

Equation (9) indicates that when $|Z_{2L}|$ is constant, the ratio A/B, which represents the ratio of the distances C_1A_1 to C_1B_1 (Fig. 9), is also constant. Hence it can be shown that the locus of C_1 is a circle, the centre of which is on line A_1B_1, and, furthermore, that all such circles are orthogonal to any circle passing through A_1 and B_1. The mesh of constant-phase and constant-magnitude circles is consequently made up of two families of mutually-orthogonal circles. A typical set of such circles is shown in Fig. 10.

The orthogonality property mentioned above can be deduced directly from the form of equation (2) which suggests that Z_{1in} and Z_{2L} are two conjugate functions and it is possible to pass from one to the other through the use of a conformal transformation.[4] The constant-phase lines and constant-magnitude circles in the load-impedance diagram are obviously two families of orthogonal circles. Hence the corresponding families of circles, in the input-impedance diagram, are also orthogonal.

When the magnitude of the load impedance



equals $|Z_{2\infty}|$, the ratio of the lengths C_1A_1 to C_1B_1 is unity. The corresponding constant-magnitude circle is in this case the straight line M_1M_2 (Fig. 9).

To construct a constant-modulus circle corresponding to a given load impedance we first determine a point C_1 (Fig. 9), such that

$$\frac{|C_1A_1|}{|C_1B_1|} = \frac{|Z_{2L}|}{|Z_{2\infty}|} \quad . . \quad . . \quad . . \quad . . \quad (12)$$

We now draw a circle through A_1 and B_1, the reactance circle C, say, and we determine point

C'_1, on line CC_1, such that $|CC_1| \times |CC'_1| = R^2$, where R is the radius of the circle of centre C. The circle of centre M, on A_1B_1, passing through C_1 and C'_1, can be shown to be orthogonal to circle C; and since C_1 satisfies equation (12), it is the desired constant-magnitude circle.

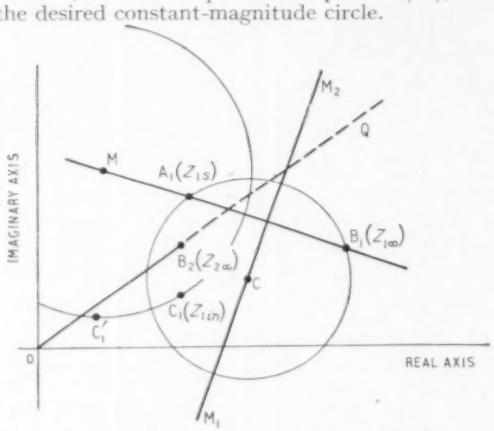


Fig. 9. *Input-impedance diagram; construction of constant-modulus circle corresponding to a load impedance of a given magnitude.*

For the particular configuration of A_1, B_1 and B_2, chosen in Fig. 9, we find that input impedances, corresponding to magnitudes of load impedances smaller than $|Z_{2\infty}|$, are represented by points to the left of M_1M_2, the perpendicular to A_1B_1, drawn through C. Similarly, input impedances corresponding to magnitudes of load impedances larger than $|Z_{2\infty}|$, are represented by points to the right of M_1M_2.

2.3. *Constant-Resistance Circles*

We shall now investigate the effect of varying the reactance component of the load impedance while leaving the resistance component constant.

It is convenient for this study to rearrange

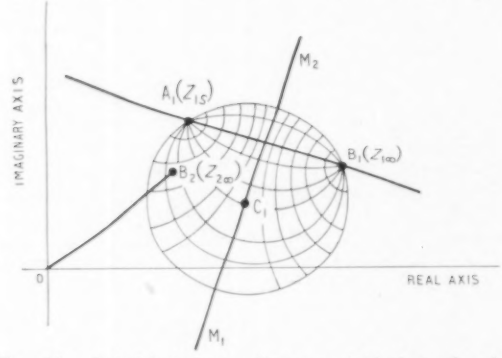


Fig. 10. *Input-impedance diagram; typical mesh of constant-phase circles (centres on M_1M_2) and constant-magnitude circles (centres on A_1B_1).*

equation (2) as follows

$$(Z_{1in} - Z_{1\infty})\ (Z_{2L} + Z_{2\infty}) = Z_{1\infty}\ (Z_{2s} - Z_{2\infty}) \quad .. \quad .. \quad .. \quad (13)$$

If OP and OB_2 represent the impedances Z_{2L} and $Z_{2\infty}$, respectively (Fig. 11), then OP′ represents the impedance $(Z_{2L} + Z_{2\infty})$. Similarly if OA_1, OB_1 and OQ represent the impedances Z_{1s}, $Z_{1\infty}$ and Z_{1in}, respectively (Fig. 12), then B_1Q represents the impedance $(Z_{1in} - Z_{1\infty})$.

According to equation (13) the product of vectors OP′ and B_1Q is equal to a complex constant, or in other words B_1Q is equal to the inverse of OP′, multiplied by the complex number

$$Z_{1\infty}(Z_{2s} - Z_{2\infty}) = Ue^{j\gamma} \quad .. \quad .. \quad (14)$$

where U and γ are real constants. Let OP′ be moved to $B_1P'_1$, in the input impedance diagram, and B_1Q_1 be a vector such that

$$B_1P'_1 \times B_1Q_1 = 1 \quad .. \quad .. \quad .. \quad (15)$$

To obtain B_1Q we rotate B_1Q_1 through the angle γ and multiply its magnitude by U. We have thus performed a complex inversion, around B_1 as pole, of inversion factor $Ue^{j\gamma}$.

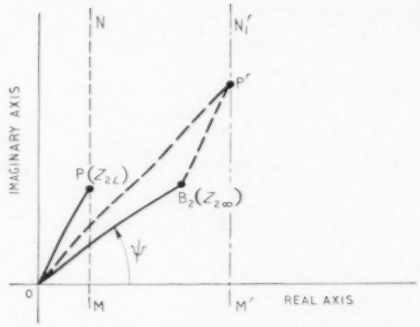


Fig. 11. *Load-impedance diagram; constant-resistance locus.*

If now the resistance component of the load impedance is kept constant, while its reactance component is varied, the locus of P′ in the load-impedance diagram is the straight line M′N′, while that of P'_1, in the input-impedance diagram, is M_1N_1. It can easily be shown, using equation (15), that the locus of Q_1 is a circle of centre D_1, on the perpendicular B_1S to M_1N_1. The subsequent transformation of Q_1 to Q, transforms circle D_1 to a circle of centre D.

For every specific value of the resistance component of the load impedance there exists a circle such as D_1 and a circle such as D. All circles such as D_1 have their centres on B_1S, therefore all circles such as D have their centres on a straight line through B_1. Since the reactance circle of centre C is also a constant-resistance circle, it follows that all constant-resistance circles have their centres on line B_1C.

It can also be shown, using the diagrams of Figs. 11 and 12, that the radius r of a constant-resistance circle is given by the formula

$$r = \frac{1}{2} \frac{|Z_{2\infty}| \cdot |Z_{1s} - Z_{1\infty}|}{R + |Z_{2\infty}| \cos \psi} \quad .. \quad .. \quad (16)$$

where R is the constant-load resistance.

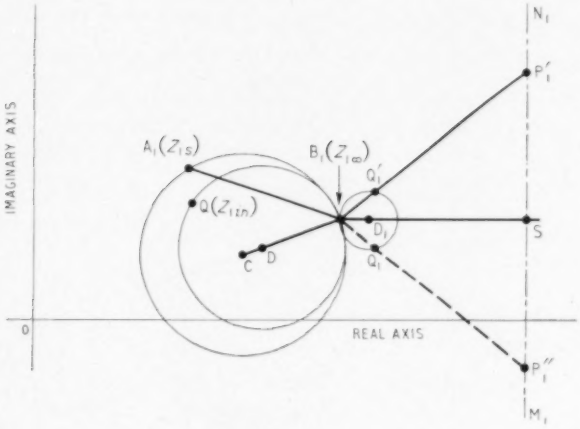
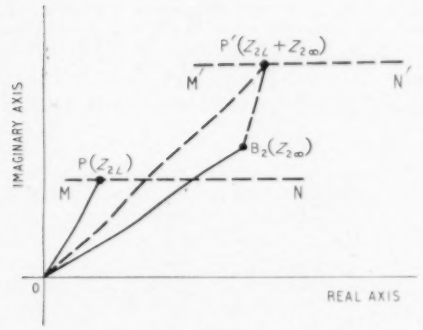


Fig. 12 (above). *Input-impedance diagram; constant-resistance circles.*

Fig. 13 (below). *Load-impedance diagram; constant-reactance locus.*

Fig. 14 (right). *Input-impedance diagram; constant-reactance circles.*



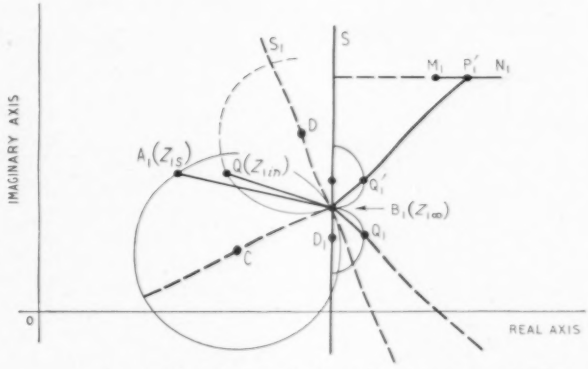
2.4. *Constant-Reactance Circles*

If the reactance of the load impedance Z_{2L} is kept constant, while its resistance is varied (Fig. 13), the locus of P', the point representing $(Z_{2\infty} + Z_{2L})$ is a straight line M'N', parallel to the real axis. If Q represents the input impedance corresponding to P (Fig. 14) then, according to equation (13), the vectors OP' and B_1Q are the complex inverse of each other. We can transfer OP' to $B_1P'_1$ and consider B_1 to be the pole of the inversion.

As in Sect. 2.3, we can obtain Q_1, the inverse of P'_1, and derive Q from Q_1 by a complex multiplication of $Z_{1\infty}(Z_{2s} - Z_{2\infty})$. Since the locus of P' is the straight line M_1N_1, the locus of Q_1 is an arc of circle, with its centre D_1 on B_1S, the line parallel to the imaginary axis, passing through B_1, and the locus of Q is therefore also an arc of circle of centre D.

B_1S is the locus of the centres of all circles such as D_1, consequently the locus of the centres of all circles such as D is a straight line B_1S_1.

Since the constant-resistance and constant-reactance lines in the load-impedance diagram are mutually orthogonal, it follows (Sect. 2.2) that the constant-resistance and constant-reactance circles in the input-impedance diagram are also mutually orthogonal. From this we deduce that the locus B_1S_1 of the centres of the constant-reactance circles is the tangent at B_1 to all the constant-resistance circles, and in particular to the reactance circle of centre C.



It can be shown, using the diagrams of Figs. 13 and 14, that the radius x of a constant-reactance circle is given by the formula

$$x = \frac{1}{2} \frac{|Z_{2\infty}||Z_{1s} - Z_{1\infty}|}{X + |Z_{2\infty}| \sin \psi} \quad .. \quad .. \quad (17)$$

where X is the constant-load reactance.

The radius x becomes infinite when

$$X = -|Z_{2\infty}| \sin \psi \quad .. \quad .. \quad .. \quad (18)$$

and the corresponding constant-reactance circle becomes the straight line B_1C. The load-impedance locus is a line parallel to the real axis and passing through B_2, while the input-impedance

locus is the diameter of the reactance circle, passing through B_1.

A typical mesh of constant-resistance and constant-reactance circles is shown in Fig. 15. The similarity with constant-resistance and

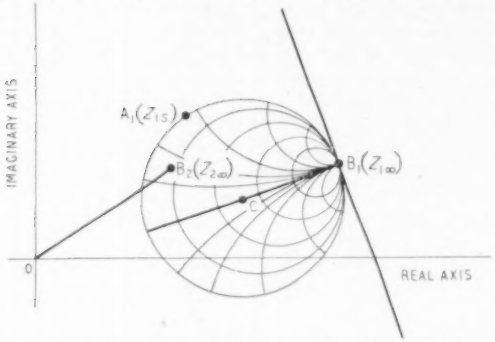


Fig. 15. *Input-impedance diagram; constant-resistance and constant-reactance circles.*

constant-reactance circles in the Smith Chart is obvious. Smith Chart markings could be used provided R is normalized to $|Z_{2c}| \cos\psi$ and X to $|Z_{2c}| \sin\psi$.

2.5. *Circles of Constant-Load Reflection Coefficient*

Another family of circles of great importance, from the practical point of view, is made up of the input-impedance loci obtained when the load impedance varies in such a way as to keep its reflection coefficient with respect to unity constant.

This reflection coefficient we have already defined in Section (1) by equation (1). The importance of this family of circles lies in the fact that a constant-reflection coefficient termination can in practice be very easily realized in high-

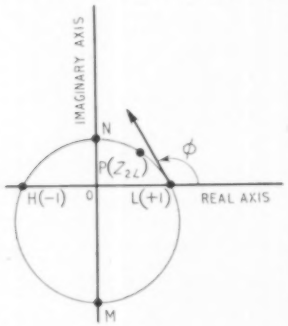


Fig. 16. *Load-impedance diagram; load-impedance locus for constant phase of reflection coefficient.*

frequency systems, and in the case of low-loss networks it is the only kind of termination which can provide reasonably accurate results.

If the locus of the load impedance Z_{2L} is a circle passing through the points L and H (Fig. 16), which represent the impedances $Z_{2L} = +1$ and $Z_{2L} = -1$, respectively, then the phase angle of the load-reflection coefficient is constant and equal to the angle ϕ. The proof of this can be very simply obtained from equation (1), using the method illustrated in Sect. 2.1.

The corresponding input-impedance locus (Fig. 17) is therefore also a circle passing through the points H_1 and L_1 corresponding to H and L, and cutting the resistance circle R (which corresponds to the real axis in the load-impedance diagram), at an angle ϕ.

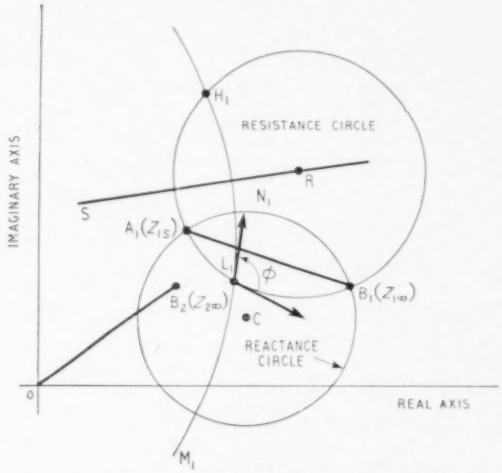
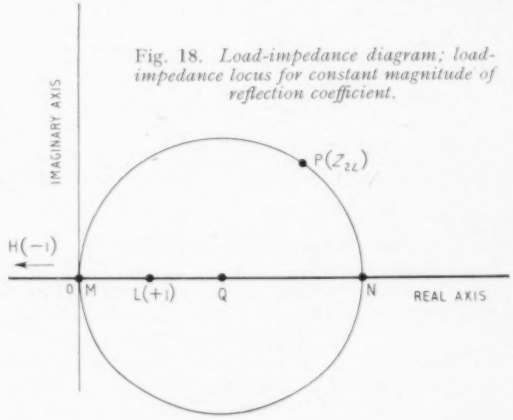


Fig. 17. *Input-impedance diagram; input-impedance locus for constant phase of reflection coefficient.*

All circles of constant phase of reflection coefficient in the load-impedance diagram pass through H and L and they are consequently orthogonal to the imaginary axis. The corresponding circles in the input-impedance diagram consequently pass through H_1 and L_1 and they are orthogonal to the reactance circle. It follows that points H_1 and L_1 are the inverse of each other about the reactance circle (i.e., if X_0 is the

Fig. 18. *Load-impedance diagram; load-impedance locus for constant magnitude of reflection coefficient.*



radius of the reactance circle, then $|CL_1| \times |CH_1| = X_0^2$).

If, on the other hand, the magnitude of the load-reflection coefficient remains constant, while the load impedance, Z_{2L}, varies, the representative point P (Fig. 18) describes a circle centred on the real axis.

This can be deduced from equation (1) if we rewrite it as follows*

$$K_{2L} = \frac{Z_{2L} - 1}{Z_{2L} + 1} = \frac{\mathbf{OP} + \mathbf{OH}}{\mathbf{OP} + \mathbf{OL}} = \frac{\mathbf{LP}}{\mathbf{HP}} \quad .. \quad (19)$$

If the magnitude of K_{2L} is constant then equation (19) shows that the ratio of the distances from P to the two fixed points L and H is constant. It follows that the locus of P is a circle centred on the straight line through L and H.

Since K and Z are, according to equation (1), conjugate functions, the circles of constant-magnitude and those of constant-phase of

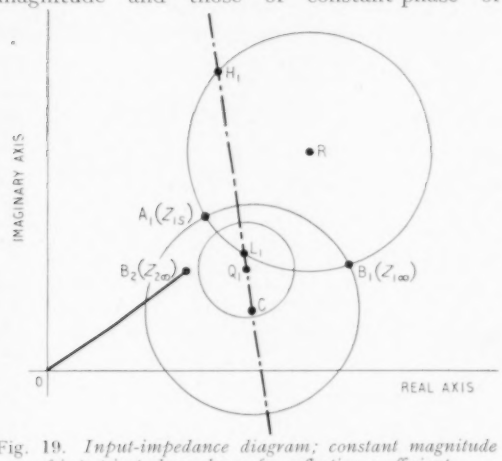


Fig. 19. *Input-impedance diagram; constant magnitude of input-impedance locus for reflection coefficient.*

reflection coefficient, both in the load-impedance and in the input-impedance diagrams, are orthogonal.

A typical mesh of constant-phase and constant-magnitude of reflection coefficient circles is shown in Fig. 20. The constant-phase circles pass through points L_1 and H_1 and therefore the constant-magnitude circles have their centres on line L_1H_1 (Fig. 19). The reactance circle is one of the latter circles, since the magnitude of the reflection coefficient K_{2L} of a purely reactive load impedance Z_{2L} is always unity; the phase of K_{2L} depends, of course, on the magnitude of Z_{2L}.

3. Transformations in the Smith Chart

Instead of plotting impedances in Cartesian co-ordinates, we can make use of the set of curvilinear co-ordinates known as the Smith Chart. The use of this new set of co-ordinates is equivalent to a change of variable from impedance Z to reflection coefficient K, according to the relation expressed by equation (1), and to a plot

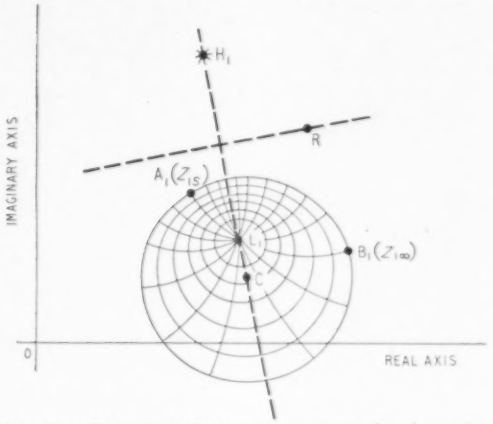


Fig. 20. *Input-impedance diagram; mesh of constant magnitude and constant phase of load-reflection coefficient circles.*

of K in Cartesian co-ordinates. To emphasize the relationship between the Smith Chart representation and a plot of reflection coefficients, we shall sometimes refer to the plot of a reflection-coefficient locus on the Smith Chart as a K-diagram, by analogy with impedance diagrams. We shall, therefore, have input K-diagrams and load K-diagrams, and the object of this section is to find the correspondence between load and input K-diagrams for four-terminal networks.

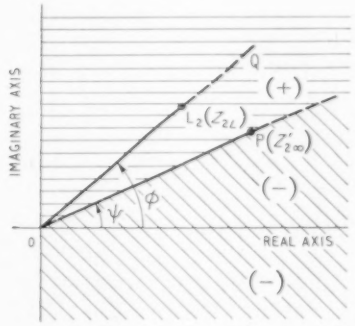


Fig. 21. *Load-impedance diagram; a constant-phase locus (line OQ).*

This correspondence can be obtained algebraically by making use of equations (1) and (5), from which we can derive the following relation

$$\frac{K_{1in} - K_{1s}}{K_{1in} - K_{1\infty}} = -\frac{Z_{2L}}{Z_{2\infty}} \frac{1 - K_{1s}}{1 - K_{1\infty}} \quad .. \quad (20)$$

*Note: The use of black-face type for a symbol such as OP denotes the vector OP.

This equation is of the same nature as equation (5) and therefore capable of the same geometrical interpretations.

3.1. *Constant-Phase Circles in the Smith Chart*

Let A_1, B_1 and C_1 (Fig. 22) be the points representing the reflection coefficients of the short-circuit input impedance (K_{1s}), the open-circuit input impedance ($K_{1\infty}$) and the input impedance (K_{1in}) for an arbitrary load impedance Z_{2L}. We can express the vectors appearing in equation (20) thus

$$K_{1\infty} - K_{1in} = |K_{1\infty} - K_{1in}|\, e^{j\phi_\infty} \quad .. \quad (21)$$

$$K_{1s} - K_{1in} = |K_{1s} - K_{1in}|\, e^{j\phi_s} \quad .. \quad (22)$$

$$\frac{1 - K_{1s}}{1 - K_{1\infty}} \cdot \frac{1}{Z_{2\infty}} = \frac{1 + Z_{1\infty}}{1 + Z_{1s}} \cdot \frac{1}{Z_{2\infty}} = \frac{1}{Pe^{j\psi}} \quad (23)$$

$$Z_{2L} = |Z_{2L}|\, e^{j\phi} \quad .. \quad .. \quad .. \quad .. \quad (24)$$

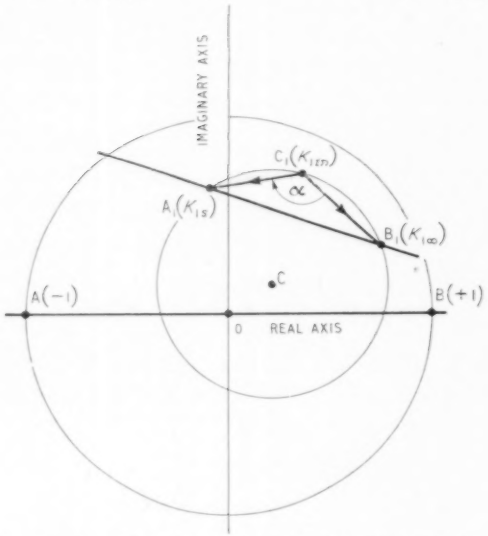


Fig. 22. *Input reflection-coefficient (K) diagram; constant-phase circle (circle of centre C).*

Equation (20) now becomes

$$\frac{|K_{1s} - K_{1in}|}{|K_{1\infty} - K_{1in}|}\, e^{j(\phi_s - \phi_\infty)} = \frac{|Z_{2L}|}{P}\, e^{j(\pi + \phi - \psi)} \quad (25)$$

and therefore $$\frac{|K_{1s} - K_{1in}|}{|K_{1\infty} - K_{1in}|} = \frac{|Z_{2L}|}{P} \quad .. \quad (26)$$

and $$\alpha = \phi_s - \phi_\infty = \pi + \phi - \psi \quad .. \quad (27)$$

where P and ψ are functions of the network alone and α is the angle $\angle B_1C_1 . C_1A_1$.

If the load impedance, Z_{2L}, varies in such a way that its phase angle, ϕ, remains constant (Fig. 21), then angle α is also constant and the locus of C_1 is an arc of circle through A_1 and B_1. It follows that for every load-impedance phase angle such as ϕ, there exists a constant-phase circle passing through A_1 and B_1.

Let OP represent the vector $Pe^{j\psi}$ in the load-impedance diagram. We can divide this diagram into two areas, meeting along OP, as indicated in Fig. 21, each labelled with the sign of the half of the imaginary axis it contains. When the load impedance is in the positive area angle ϕ is greater than ψ, consequently α is larger than π radians and C_1 must lie above the line A_1B_1, for the particular disposition of A_1 and B_1, chosen in the figure. Similarly, if the load impedance is in the negative area, then C_1 lies below A_1B_1.

If the load impedance is a pure reactance, the locus of the point C_1 is a complete circle of centre C. For any other phase angle, the corresponding constant-phase locus is an arc of circle wholly contained within the reactance circle. As in the case of the input-impedance mapping, the network transforms the right-hand side of the load-impedance diagram into the interior of the reactance circle, plotted in the input K-diagram.

Let us define an impedance $Z'_{2\infty}$ such that

$$Z'_{2\infty} = Z_{2\infty} \frac{1 + Z_{1s}}{1 + Z_{1\infty}} = Z_{2\infty} \frac{1 - K_{1\infty}}{1 - K_{1s}} \quad (28)$$

In terms of $Z'_{2\infty}$ equation (20) becomes

$$\frac{K_{1in} - K_{1s}}{K_{1in} - K_{1\infty}} = -\frac{Z_{2L}}{Z'_{2\infty}} \quad .. \quad .. \quad (29)$$

and, but for a change in one of the parameters, equations (29) and (5) are identical. It follows that the geometrical constructions established in Section 2, hold equally well for the input K-diagrams, provided the input parameters are

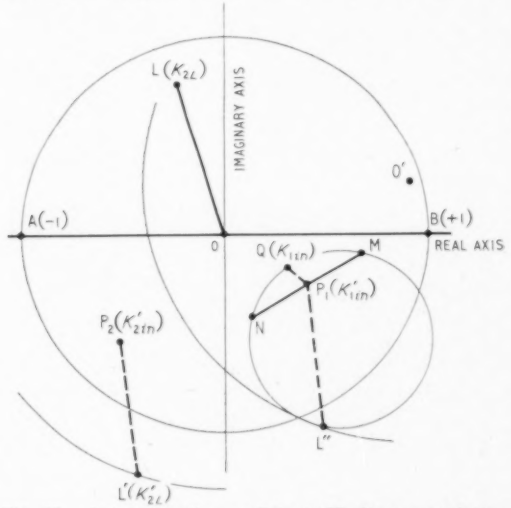


Fig. 23. *Input reflection-coefficient (K) diagram; relation between complex insertion loss and input and load reflection coefficients, for unsymmetrical networks.*

replaced by their reflection coefficients and that $Z_{2\infty}$ is replaced by $Z'_{2\infty}$, the other output parameter remaining unchanged.

In view of this, the derivation of constant-phase, constant-modulus, constant-resistance and constant-reactance in the input K-diagram will not be repeated. The constant reflection-coefficient circles, however, because of their great importance, will be examined more thoroughly, at the risk of a little repetition, in the next section.

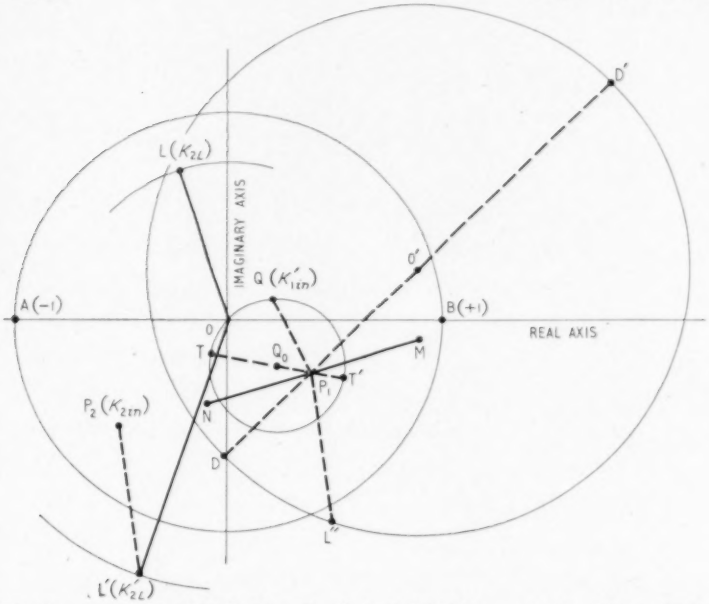


Fig. 24. *Input reflection-coefficient diagram; the derivation of a constant-load reflection-coefficient circle.*

3.2. *Constant-Load Reflection-Coefficient Circles*

The importance of this family of circles is due to the fact that a simple determination of the insertion loss of a quadripole is made possible through their use.

It can be proved that the complex insertion loss S of a quadripole is related to the network and load parameters by the following equation

$$S^2 = (K_{1in} - K'_{1in})(K'_{2L} - K'_{2in}) \quad .. \quad (30)$$

where

S = Complex insertion loss of four-terminal network defined as the ratio of the currents flowing into a load impedance Z_{2L} when the latter is (a) connected direct to a generator (assumed here to be of unity impedance) and (b) connected to the same generator through the network.

K'_{1in} = Reflection coefficient at terminals 1 of network when the load impedance is unity.

K_{1in} = Reflection coefficient at terminals 1 of network when the load impedance has any value Z_{2L}.

K'_{2in} = Reflection coefficient at terminals 2 of the network when the load impedance is equal to unity.

K_{2L} = Reflection coefficient of Z_{2L}, the load impedance at terminals 2.

K'_{2L} = The inverse of K_{2L} (i.e., $K'_{2L} = 1/K_{2L}$).

(It is important to note in connection with the above definitions that the assumption of a generator impedance of unity implies that impedances are normalized to the generator impedance. The reflection coefficients are also expressed in terms of the same impedance.)

Let P_1 and P_2 (Fig. 23) represent K'_{1in} and K'_{2in}, while L and Q represent K_{2L} and K_{1in}. Given L we can determine the point L', representing K'_{2L}, the inverse of K_{2L}.

We have

$$\mathbf{P_1Q} = K_{1in} - K'_{1in}$$
$$\mathbf{P_2L'} = K'_{2L} - K'_{2in}.$$

If L'' is a point such that $\mathbf{P_1L''} = \mathbf{P_2L'}$ and M and N such that $\mathbf{PM} = -\mathbf{PN} = S$, then we can rewrite equation (30) as

$$\mathbf{PN}^2 = \mathbf{PM}^2 = \mathbf{P_1Q}.\mathbf{P_1L''} \quad .. \quad .. \quad (31)$$

which is the well-known condition for the four-points M, N, Q and L'' to lie on a common circle. **PM** and **PN** are the geometric means of $\mathbf{P_1Q}$ and $\mathbf{P_1L''}$ and it can be easily proved that the direction NP_1M bisects the angle QP_1L''.

If now the load impedance Z_{2L} varies in such a way that the magnitude of its reflection coefficient remains constant, the locus of L (Fig. 24) is a circle, concentric with and within the unity circle, while the locus of L' is a circle concentric with the unity circle, but outside it. The locus of L'' is the displaced locus of L'; its centre is at O', such that $\mathbf{OO'} = \mathbf{P_2P_1}$. According to equation (31) $\mathbf{P_1Q}$ can be obtained by multiplying the inverse of $\mathbf{P_1L''}$ by the constant S^2. Since the inverse of a circle is also a circle, it follows that the locus of Q is a circle, the centre of which is on TT', the straight line through P_1, such

that NP_1M is the bisector of angle $D'P_1T'$, where DD' is the diameter through P_1 of the locus of L″.

For every magnitude $|K_{2L}|$ of load reflection coefficient, there corresponds a circle for the input reflection coefficient K_{1in}. These circles, which we shall call, for brevity, the Q circles,

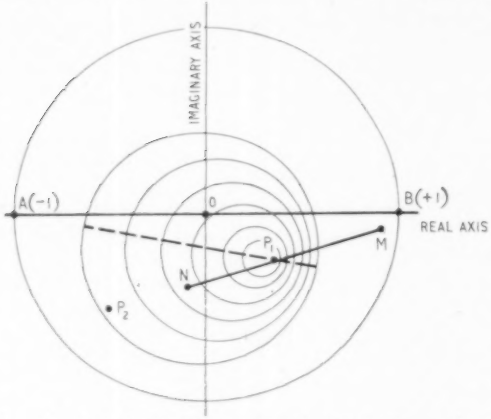


Fig. 25. *Input reflection-coefficient plane; the mesh of constant-load reflection-coefficient circles.*

are all centred on line TT′. When the magnitude of K_{2L} is very small (i.e., the load impedance is nearly equal to unity) the locus of L is a very small circle, that of L″ is a very large one and the corresponding Q circle is very small. Its radius becomes vanishingly small as $|K_{2L}|$ tends to zero. When $|K_{2L}|$ is at its maximum value of unity, the corresponding Q circle coincides with the reactance circle. A typical mesh of Q circles is shown in Fig. 25.

It is useful to note that in the case of symmetrical networks, since P_1 and P_2 are one and the same point, the centre O′ of the displaced load circles coincides with O.

We can now calculate the radius of a given Q circle from Fig. 24 and equation (31). We have

$$|P_1T'| = \frac{|S^2|}{|P_1D'|} \quad .. \quad .. \quad .. \quad .. \quad (32)$$

$$|P_1T| = \frac{|S^2|}{|P_1D|} \quad .. \quad .. \quad .. \quad .. \quad (33)$$

$$|P_1D'| = |K'_{2in}| + |K'_{2L}| \quad .. \quad .. \quad (34)$$

$$|P_1D| = |K'_{2L}| - |K'_{2in}| \quad .. \quad .. \quad (35)$$

If R is the radius of a given constant-reflection coefficient circle, we have

$$2R = |P_1T| + |P_1T'| \quad .. \quad .. \quad .. \quad (36)$$

$$2R = |S^2|\left[\frac{1}{|K'_{2L}| - |K'_{2in}|} + \frac{1}{|K'_{2L}| + |K'_{2in}|}\right] \quad .. \quad .. \quad .. \quad (37)$$

$$2R = |S^2|\frac{2K'_{2L}}{|K'_{2L}|^2 - |K'_{2in}|^2} \quad .. \quad .. \quad (38)$$

Replacing K'_{2L} by $1/K_{2L}$ we obtain

$$R = \frac{|S^2|\,|K_{2L}|}{1 - |K'_{2in}|^2\,|K_{2L}|^2} \quad .. \quad .. \quad (39)$$

(To be continued)

L.F. COMPENSATION FOR VIDEO AMPLIFIERS

Part 2. Negative-Feedback Amplifiers

By J. E. Flood, Ph.D., A.M.I.E.E.

SUMMARY.—The method outlined in Part 1 for designing amplifiers with indicial responses having the highest possible order of compensation is applied to negative-feedback amplifiers. If β is independent of frequency, the overall response is an optimum when the indicial response of the μ-path is an optimum and the order of compensation is the same. If β is made frequency-dependent by including a capacitor in the feedback path, compensation can be obtained which is one order higher than that obtainable from the μ-path alone.

The method is applied to some simple circuits, including cases where a coupling or a decoupling circuit is outside the feedback loop. Some of the results are confirmed by experiment.

1. Introduction

DISTORTION at low frequencies affects the indicial response of an amplifier (i.e., its response to a unit step) causing it to depart from the ideal flat top. The indicial response can be made as flat as possible by causing as many as possible of its differential coefficients to be zero at $t = 0$. If the first r derivatives vanish, the indicial response is said to have rth-order compensation. A general method for obtaining the highest possible order of compensation was described in Part 1 of this paper. In this part of the paper the method is applied to some typical negative-feedback amplifier circuits.

2. Amplifier with β Independent of Frequency

If the indicial response of the μ-path of the amplifier is $\mu(p)\mathbf{1}$ and the feedback fraction β_0 is independent of frequency, then the overall indicial response of the amplifier with feedback is

$$h(p)\,\mathbf{1} = \frac{\mu(p)}{1+\beta_0\mu(p)}\,\mathbf{1}$$

The sign convention is chosen so that μ and β_0 are both positive quantities when the feedback is negative.

Let
$$\mu(p) = \mu_0 \frac{p^n + a_1p^{n-1} + a_2p^{n-2} + \ldots + a_qp^{n-q} + \ldots + a_n}{p^n + b_1p^{n-1} + b_2p^{n-2} + \ldots + b_qp^{n-q} + \ldots + b_n} \quad \ldots \quad (1)$$

Then
$$h(p)\,\mathbf{1} = \frac{\mu_0}{1+\mu_0\beta_0} \cdot \frac{p^n + a_1p^{n-1} + \ldots + a_qp^{n-q} + \ldots + a_n}{p^n + d_1p^{n-1} + \ldots + d_qp^{n-q} + \ldots + d_n}\,\mathbf{1} \quad \ldots \quad (2)$$

where $d_q = [b_q + \mu_0\beta_0 a_q]/[1+\mu_0\beta_0]$ .. (3)

The overall response therefore has rth-order compensation when $a_1 = d_1$, $a_2 = d_2$, $\ldots . a_r = d_r$. But from equation (3), $a_q = d_q$ when $a_q = b_q$. The overall response of the feedback amplifier therefore has rth-order compensation when the response of its μ-path has rth-order compensation. In particular, a maximally-flat overall response ($r = n-1$) requires that the μ-path shall have a maximally-flat indicial response. The procedure for obtaining an optimum overall response is thus to design the μ-path to have an indicial response with the highest possible order of compensation.

3. Effect of Networks outside the Feedback Loop

3.1. *Coupling Circuit outside the Feedback Loop*

If the feedback voltage is applied to the cathode of the first valve, as shown in Fig. 1, then any coupling circuit between its grid and the input terminals is outside the feedback loop. If the coupling circuit comprises capacitance C_c and resistance R_g, its indicial response is $p/(p+\alpha)\mathbf{1}$ where $\alpha = 1/C_cR_g$. If the indicial response of the amplifier is given by equation (2), then the overall response, including the coupling circuit, is given by

$$h(p)\,\mathbf{1} = \frac{p}{p+\alpha} \cdot \frac{\mu_0}{1+\mu_0\beta_0} \cdot \frac{p^n + a_1p^{n-1} + \ldots + a_qp^{n-q} + \ldots + a_n}{p^n + d_1p^{n-1} + \ldots + d_qp^{n-q} + \ldots + d_n}\,\mathbf{1}$$

$$= \frac{\mu_0}{1+\mu_0\beta_0} \cdot \frac{p^{n+1} + a_1p^n + \ldots + a_qp^{n-q+1} + \ldots + a_np}{p^{n+1} + D_1p^n + \ldots + D_qp^{n-q+1} \ldots + D_np + D_{n+1}}\,\mathbf{1} \quad \ldots \quad (4)$$

where $D_q = d_q + \alpha\, d_{q-1}$ .. .. (5)

$\therefore\ (1+\mu_0\beta_0)\, D_q = b_q + \alpha\, b_{q-1} + \mu_0\beta_0(a_q + \alpha a_{q-1})$ .. .. (6)

from equation (3)

$\therefore\ a_q = D_q$ when

$$a_q = b_q + \alpha\,(b_{q-1} + \mu_0\beta_0\, a_{q-1}) \quad \ldots \quad (7)$$

The overall response now has rth-order compensation when the coefficients $a_1, a_2 \ldots a_r$ and $b_1, b_2 \ldots b_r$ of the indicial response of its μ-path satisfy equation (7).

It is interesting to compare the effect of adding a coupling circuit to the amplifier outside its feedback loop with the effect of adding a coupling circuit inside the loop. If a coupling circuit with time-constant $1/\alpha'$ is added to the amplifier whose indicial response is given by equation (1), the response of the μ-path including the coupling circuit is

$$\mu(p)\,\mathbf{1} = \mu_0 \frac{p}{p+\alpha'} \cdot \frac{p^n + a_1p^{n-1} + \ldots + a_qp^{n-q} + \ldots + a_n}{p^n + b_1p^{n-1} + \ldots + b_qp^{n-q} + \ldots + b_n}\,\mathbf{1}$$

$$= \mu_0 \frac{p^{n+1} + a_1p^n + \ldots + a_qp^{n-p+1} + \ldots + a_np}{p^{n+1} + B_1p^n + \ldots + B_qp^{n-q+1} + \ldots + B_np + B_{n+1}}\,\mathbf{1}$$

where $B_q = b_q + \alpha'\, b_{q-1}$

The overall indicial response is therefore given by equation (4) where, from equation (3):

$$D_q = [B_q + \mu_0\beta_0\, a_q] \,/\, [1+\mu_0\beta_0]$$

$$\therefore\ (1+\mu_0\beta_0)\, D_q = b_q + \alpha'\, b_{q-1} + \mu_0\beta_0\, a_q \quad (8)$$

and $a_q = D_q$ when

$$a_q = b_q + \alpha'\, b_{q-1} \quad \ldots \quad (9)$$

The overall response now has rth-order compensation when the coefficients $a_1, a_2 \ldots a_r$ and $b_1, b_2 \ldots b_r$ satisfy equation (9). The difference in form between equation (6) and equation (8) shows that, in general, the addition of a coupling circuit has a different effect on the overall indicial response depending on whether it is inside or outside the feedback loop.

3.2. *Anode Decoupling Circuit outside Feedback Loop*

If the feedback voltage is taken from the cathode of the last valve, as shown in Fig. 1, then any decoupling network in its anode circuit is

outside the feedback loop. If the last stage has an anode load resistance R_a, a decoupling resistance R_d and decoupling capacitance C_d, its indicial response is of the form

$$G\frac{p+(\gamma+\lambda)}{p+\lambda}\,1$$

where G is the nominal stage gain, $\gamma = 1/R_aC_d$ and $\lambda = 1/R_dC_d$.

If the indicial response of the amplifier, excluding the last stage, is given by equation (2), then the overall response including the last stage, with its decoupling circuit, is

$$h(p)\,1 = \frac{p+(\gamma+\lambda)}{p+\lambda}\cdot\frac{\mu_0}{1+\mu_0\beta_0}\cdot\frac{p^n+a_1p^{n-1}+\ldots+a_qp^{n-q}+\ldots+a_n}{p^n+d_1p^{n-1}+\ldots+d_qp^{n-q}+\ldots+d_n}\,1$$
$$= \frac{\mu_0}{1+\mu_0\beta_0}\cdot\frac{p^{n+1}+C_1p^n+\ldots+C_qp^{n-q+1}+\ldots+C_{n+1}}{p^{n+1}+D_1p^n+\ldots+D_qp^{n-q+1}+\ldots+D_{n+1}}\,1$$

where $C_q = a_q + (\gamma+\lambda)\,a_{q-1}$

and $D_q = d_q + \lambda d_{q-1}$

$$\therefore\ (1+\mu_0\beta_0)\,D_q = b_q + \lambda b_{q-1} + \mu_0\beta_0(a_q + \lambda a_{q-1}) \quad .. \quad .. \quad (10)$$

and $C_q = D_q$ when

$$a_q + a_{q-1}[\lambda + \gamma(1+\mu_0\beta_0)] = b_q + b_{q-1} \quad (11)$$

The overall indicial response therefore has rth-order compensation when the coefficients $a_1, a_2 \ldots a_r$ and $b_1, b_2 \ldots b_r$ of the indicial response of its μ-path satisfy equation (11).

It is interesting to compare the effect of adding a decoupling circuit to the amplifier outside its feedback loop with the effect of adding a decoupling circuit inside the loop. If a decoupling circuit with time-constants $1/\gamma'$ and $1/\lambda'$ is added to the amplifier whose indicial response is given by equation (1), the response of the μ-path including the decoupling circuit is

$$\mu(p)\,1 = \mu_0\frac{p+(\gamma'+\lambda')}{p+\lambda'}\cdot\frac{p^n+a_1p^{n-1}+\ldots+a_qp^{n-q}+\ldots+a_n}{p^n+b_1p^{n-1}+\ldots+b_qp^{n-q}+\ldots+b_n}\,1$$
$$= \mu_0\frac{p^{n+1}+A_1p^n+\ldots+A_qp^{n-q+1}+\ldots+A_{n+1}}{p^{n+1}+B_1p^n+\ldots+B_qp^{n-q+1}+\ldots+B_{n+1}}\,1$$

The overall indicial response is therefore given by

$$h(p)\,1 = \frac{\mu_0}{1+\mu_0\beta_0}\cdot\frac{p^{n+1}+A_1p^n+\ldots+A_qp^{n-q+1}+\ldots+A_{n+1}}{p^{n+1}+D_1p^n+\ldots+D_qp^{n-q+1}+\ldots+D_{n+1}}\,1$$

where $A_q = a_q + (\gamma'+\lambda')\,a_{q-1}$

and $B_q = b_q + \lambda' b_{q-1}$

where, from equation (3),

$$D_q = [B_q + \mu_0\beta_0 A_q]/(1+\mu_0\beta_0)$$
$$\therefore\ (1+\mu_0\beta_0)\,D_q = b_q + \lambda' b_{q-1} + \mu_0\beta_0[a_q + (\gamma'+\lambda')\,a_{q-1}] \quad .. \quad (13)$$

and $A_q = D_q$ when

$$a_q + a_{q-1}(\gamma'+\lambda') = b_q + b_{q-1}\lambda' \quad .. \quad (14)$$

The difference in form between equation (13) and equation (10) shows that, in general, a decoupling circuit has a different effect on the indicial response depending on whether it is inside or outside the feedback loop. However, equations (14) and (11) are identical when

$$\lambda' = \lambda \quad .. \quad .. \quad .. \quad (15)$$

and

$$\gamma' = \gamma(1+\mu_0\beta_0) \quad .. \quad (16)$$

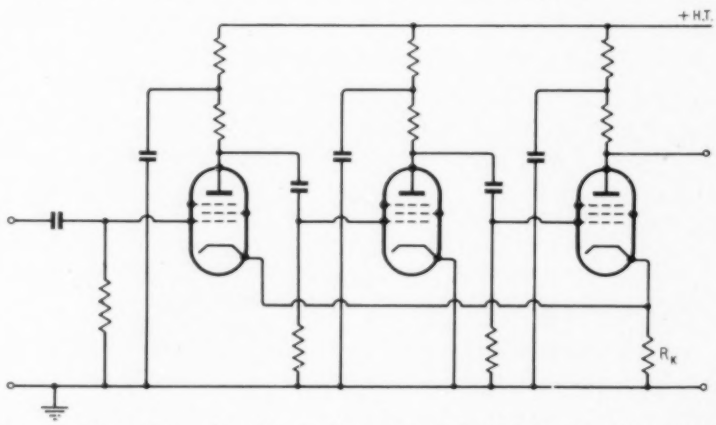


Fig. 1. *Amplifier with feedback taken from the cathode of the third valve.*

The condition for an amplifier with a decoupling circuit outside the feedback loop to have rth-order compensation is therefore the same as that for an amplifier with an additional decoupling circuit inside the loop having the same time-constant C_dR_d but $1/(1 + \mu_0\beta_0)$ times the time-constant C_dR_a.

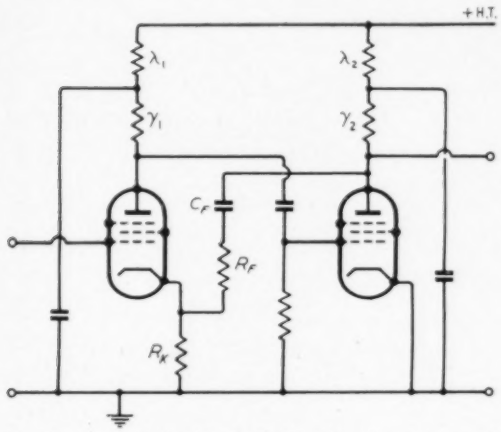


Fig. 2. *Amplifier with feedback taken from the anode of the second valve.*

4. Amplifier with a Capacitor in the Feedback Path

The amplifiers considered in the previous sections each have a feedback fraction β which is independent of frequency. However, it is often necessary to include a capacitor in the feedback path in order to block d.c., as shown in Fig. 2. β is then frequency-dependent and the indicial response of the feedback path is

$$\beta(p)\,\mathbf{1} = \beta_0 \frac{p}{p + \sigma}\,\mathbf{1} \qquad .. \qquad .. \qquad (17)$$

where $\beta_0 = R_K/(R_K + R_F)$ and $\sigma = 1/C_F(R_K + R_F)$

If the response of the μ-path is given by equation (1) then the overall indicial response of the amplifier is

$$h(p)\,\mathbf{1} = \frac{\mu(p)}{1 + \mu(p).\beta(p)}\,\mathbf{1}$$

$$= \frac{\mu_0}{1 + \mu_0\beta_0} \cdot \frac{p^{n+1} + c_1p^n + \ldots + c_qp^{n-q+1} + \ldots + c_{n+1}}{p^{n+1} + d_1p^n + \ldots + d_qp^{n-q+1} + \ldots + d_{n+1}}\,\mathbf{1}$$

where

$$c_1 = a_1 + \sigma, \qquad d_1 = (\mu_0\beta_0a_1 + b_1 + \sigma)/(1 + \mu_0\beta_0)$$

$$c_2 = a_2 + a_1\sigma, \qquad d_2 = (\mu_0\beta_0a_2 + b_2 + b_1\sigma)/(1 + \mu_0\beta_0)$$

- - - -

5. Amplifiers whose Response is Determined by Coupling Circuits Alone

5.1. *Single-Stage Amplifier*

For a single stage with a coupling circuit comprising capacitance C_c and resistance R_g

$$\mu(p) = \mu_0 \frac{R_g}{R_g + 1/pC_c} = \mu_0 \frac{p}{p + \alpha}$$

where $\alpha = 1/C_c R_g$

If β is independent of frequency then the overall indicial response is given by

$$h(p)\,1 = \frac{\mu_0 p/(p+\alpha)}{1 + \mu_0\beta_0 p/(p+\alpha)}\,1$$

$$= \frac{\mu_0}{1+\mu_0\beta_0} \cdot \frac{p}{p+\alpha'}\,1$$

where $\alpha' = \alpha/(1+\mu_0\beta_0)$

The indicial response of the feedback amplifier is thus of the same form as that of the amplifier without feedback. The effect of the feedback is to multiply the time-constant of the amplifier by the feedback factor $(1+\mu_0\beta_0)$. The initial rate at which the indicial response decays is thus reduced in the same proportion as the gain is reduced by the feedback.

If the feedback path contains a capacitor then $\beta(p)$ is given by equation (17) and

$$h(p)\,1 = \frac{\mu_0 p/(p+\alpha)}{1 + \mu_0\beta_0 p^2/(p+\alpha)(p+\sigma)}\,1$$

$$= \frac{\mu_0}{1+\mu_0\beta_0} \cdot \frac{p^2+\sigma p}{p^2 + d_1 p + d_2}\,1$$

where $d_1 = (\alpha+\sigma)/(1+\mu_0\beta_0)$

and $d_2 = \alpha\sigma/(1+\mu_0\beta_0)$

It is thus possible to obtain first-order compensation. We require $\sigma = d_1$ and this is obtained when

$$\alpha = \mu_0\beta_0\sigma \quad .. \quad .. \quad .. \quad .. \quad (21)$$

5.2. *Multi-Stage Amplifier*

For an n-stage amplifier with time-constants $1/\alpha_1, 1/\alpha_2, \ldots 1/\alpha_n$

$$\mu(p) = \mu_0 \frac{p^n}{(p+\alpha_1)(p+\alpha_2)\ldots(p+\alpha_n)}$$

The μ-path thus has a zero-order indicial response. If β is independent of frequency then the overall response also has zero-order compensation. If the feedback path contains a capacitor, then $\beta(p)$ is given by equation (17) and the overall indicial response is

$$\frac{\mu(p)}{1+\mu(p).\beta(p)}\,1 = \frac{\mu_0 p^{n+1} + \mu_0\sigma p^n}{\mu_0\beta_0 p^{n+1} + (p+\alpha_1)(p+\alpha_2)\ldots(p+\alpha_n)}\,1$$

$$= \frac{\mu_0}{1+\mu_0\beta_0} \cdot \frac{p^{n+1} + \sigma p^n}{p^{n+1} + d_1 p^n + \ldots + d_r p^{n-r+1} + \ldots + d_{n+1}}\,1 \quad .. \quad .. \quad (22)$$

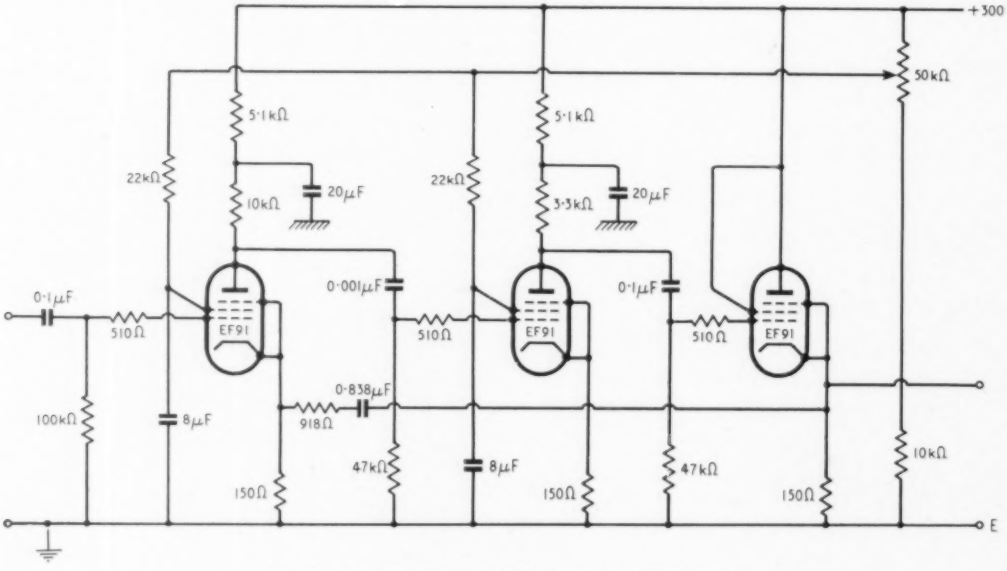


Fig. 3. *Experimental amplifier with negative feedback.*

where $d_1 = \left\{ \sigma + \sum_{r=1}^{n} \alpha_r \right\} \Big/ [1 + \mu_0\beta_0]$

First-order compensation is thus obtained when $\sigma = d_1$

$$\therefore \qquad \mu_0\beta_0\sigma = \sum_{r=1}^{n} \alpha_r \qquad .. \qquad .. \qquad .. \qquad (23)$$

If the amplifier has a resistance-capacitance coupling circuit which is outside the feedback loop then the overall indicial response is given by

$$h(p)\,\mathbf{1} = \frac{p}{p + \alpha_0} \cdot \frac{\mu(p)}{1 + \mu(p).\beta(p)}\,\mathbf{1}$$

$$= \frac{\mu_0}{1 + \mu_0\beta_0} \cdot \frac{p^{n+2} + \sigma p^{n+1}}{p^{n+2} + D_1 p^{n+1} + \ldots + D_r p^{n-r+2} + \ldots + D_{n+2}}\,\mathbf{1}$$

where $D_r = d_r + \alpha_0 d_{r-1}$ [from equations (5) and (22)].

First-order compenstion is thus obtained when

$$\sigma = D_1 = b_1 + \alpha_0$$

$$\therefore \qquad \mu_0\beta_0\sigma = \sum_{r=1}^{n} d_r + (1 + \mu_0\beta_0)\alpha_0 \qquad .. \qquad (24)$$

Comparing equations (23) and (24) it is seen that, in this case, the effect of the external coupling circuit on the condition for first-order compensation is equivalent to an additional coupling circuit inside the feedback loop with a time constant equal to that of the external coupling circuit divided by $(1 + \mu_0\beta_0)$.

6. Amplifier with a Single Stage of Coupling and Decoupling

If the amplifier has a single stage of anode decoupling and a single coupling circuit, then from equation (12) of Part 1:

$$\mu(p) = \mu_0 \frac{p^2 + p(\gamma_1 + \lambda_1)}{p^2 + p(\alpha + \lambda_1) + \alpha\lambda_1}$$

where $\alpha = 1/R_gC_c$, $\gamma_1 = 1/R_aC_d$, $\lambda_1 = 1/R_dC_d$.

The μ-path thus gives a first-order indicial response when $\gamma_1 = \alpha$. If β is independent of frequency then the overall response also has first-order compensation.

If the feedback path contains a capacitor and has time-constant $1/\sigma$ then Section 4 shows that its effect is equivalent to a second stage of anode decoupling whose time-constants $1/\gamma_2$ and $1/\lambda_2$ are given by equations (19) and (20) as

$$\gamma_2 = \mu_0\beta_0\sigma \text{ and } \lambda_2 = \sigma$$

The conditions for compensation of the feedback amplifier therefore correspond to those for an amplifier without feedback having one coupling circuit and two stages of anode decoupling. It was shown in Section 4.1 of Part 1 that this could be made to give second-order compensation. The conditions for second-order compensation, from equations (17) and (18) of Part 1 are

$$\alpha = \gamma_1 + \gamma_2$$

$$\text{and} \qquad \gamma_1\gamma_2 = \gamma_1\lambda_1 + \gamma_2\lambda_2$$

$$\text{which gives } \alpha = \gamma_1 + \mu_0\beta_0\sigma \qquad .. \qquad .. \qquad (25)$$

$$\text{and} \qquad \lambda_1 = \mu_0\beta_0\sigma(1 - \sigma/\gamma_1) \qquad .. \qquad .. \qquad (26)$$

7. Amplifier with Two Stages of Anode Coupling, One Coupling Circuit and a Capacitor in the Feedback Path

The typical feedback amplifier circuit shown in Fig. 2 has two stages of anode decoupling and one coupling circuit in the μ-path and a capacitor in the β-path. Let the two anode circuits have time-constants $1/\lambda_1$, $1/\lambda_2$, $1/\gamma_1$, $1/\gamma_2$ and the feedback path a time-constant $1/\sigma$. It is shown in Section 4 that the effect of the feedback capacitor is equivalent to a third stage of anode decoupling whose time-constants $(1/\gamma_3,\ 1/\lambda_3)$ are given by equations (19) and (20) as

$$\gamma_3 = \mu_0\beta_0\sigma$$
$$\lambda_3 = \sigma$$

The overall indicial response is therefore equivalent to that of an amplifier without feedback having one coupling circuit and three stages of anode decoupling. It was shown in Section 4.4 of Part 1 that this gives third-order compensation when the following conditions are satisfied:

$$\alpha = \gamma_1 + \gamma_2 + \gamma_3$$

$$\gamma_1\gamma_2 + \gamma_2\gamma_3 + \gamma_3\gamma_1 = \gamma_1\lambda_1 + \gamma_2\lambda_2 + \gamma_3\lambda_3$$

$$\gamma_1\lambda_1(\lambda_2 + \lambda_3) + \gamma_2\lambda_2(\lambda_1 + \lambda_3) + \gamma_3\lambda_3(\lambda_1 + \lambda_2) = \lambda_1\gamma_2\gamma_3 + \lambda_2\gamma_1\gamma_3 + \lambda_3\gamma_1\gamma_2 + \gamma_1\gamma_2\gamma_3$$

If we put $\gamma_1 = \gamma_2 = \gamma$ and $\lambda_1 = \lambda_2 = \lambda$ these equations give

$$\alpha = 2\gamma + \gamma_3 \qquad .. \qquad .. \qquad .. \qquad (27)$$

$$\lambda = \tfrac{1}{2}(\gamma + 2\gamma_3 - \gamma_3\lambda_3/\gamma) \qquad .. \qquad (28)$$

and

$$\gamma^4 + 4\gamma^2\gamma_3\lambda_3 + 2\gamma\gamma_3\lambda_3(\gamma_3 - \lambda_3) - \gamma_3{}^2\lambda_3{}^2 = 0 \qquad .. \qquad .. \qquad .. \qquad (29)$$

Substituting for γ_3 and λ_3 in equation (29) gives

$$\gamma^4 + 4\gamma^2\sigma^2\mu_0\beta_0 + 2\gamma\sigma^3\mu_0\beta_0(\mu_0\beta_0 - 1) - \sigma^4(\mu_0\beta_0)^2 = 0 \qquad .. \qquad .. \qquad (30)$$

which enables the required value of σ to be obtained for any particular numerical values of γ and $\mu_0\beta_0$. By solving equation (30) as a quadratic for $\mu_0\beta_0$ it can be shown that $\mu_0\beta_0$ is only real and positive for $0 < \gamma < \frac{1}{2}\sigma$. In most

cases $\mu_0\beta_0 \gg 1$ so that $\gamma \ll \mu_0\beta_0\sigma$ and equation (30) reduces to

$$2\gamma\sigma^3(\mu_0\beta_0)^2 - \sigma^4(\mu_0\beta_0)^2 = 0$$

$$\therefore \quad \sigma = 2\gamma \quad .. \quad .. \quad (31)$$

and substituting in equations (27) and (28) gives

$$\alpha = 2\gamma(1 + \mu_0\beta_0) \quad .. \quad .. \quad .. \quad (32)$$

and $$\lambda = \tfrac{1}{2}\gamma \quad .. \quad .. \quad .. \quad .. \quad (33)$$

The error involved in using equation (31) instead of the exact equation (30) is less than 1% for $\mu_0\beta_0 > 2$ and less than 0·1% for $\mu_0\beta_0 > 7{\cdot}4$.

8. More Complicated Circuits

Circuits which are more complicated than those considered in the previous sections can often be made to have indicial responses with higher orders of compensation, but the calculations are correspondingly more difficult. In many practical cases, however, it is sufficient to obtain first- or second-order compensation. A complicated circuit can then be dealt with by considering it as a number of simpler stages connected in tandem, each of which can be made to have the required order of compensation.

Consider, for example, an amplifier whose response is determined by two coupling circuits and two stages of anode decoupling in the μ-path, a capacitor in the β-path and one coupling circuit and one decoupling circuit outside the feedback loop.* Section 4 shows that the effect of the capacitor in the β-path on the order of compensation is equivalent to adding a decoupling circuit outside the loop. The feedback amplifier can therefore be considered as equivalent to an amplifier with two coupling circuits and two anode decoupling circuits and with β independent of frequency, connected in tandem with another amplifier without feedback which has one coupling circuit and two decoupling circuits. Section 2 shows that the feedback amplifier gives the same order of compensation as its μ-path. The two equivalent amplifiers each with two decoupling circuits can be made to have second-order compensation as shown in Section 4 of Part 1. The overall indicial response then also has second-order compensation.

9. Stability Considerations

An indicial response with a high order of compensation is very flat in the neighbourhood of $t = 0$, but that condition gives no control of the ultimate behaviour of the response with increasing time. If an amplifier has been designed to have an indicial response with a high order of compensation it does not therefore necessarily follow that it will be stable. Consequently, it is still necessary to check the $\mu\beta$ characteristics of the amplifier to ensure that it has an adequate stability margin.

Consider, for example, an amplifier whose response is entirely determined by three coupling circuits in the μ-path with equal time-constants. It can be shown that this amplifier only satisfies the Nyquist stability criterion if $\mu_0\beta_0 < 8$. If a capacitor is added to the β-path, to give first-order compensation of the indicial response, the increase in phase-shift round the feedback loop results in the amplifier being unstable if $\mu_0\beta_0 > 4{\cdot}05$. If more feedback is required, the time-constants of the coupling circuits must be made unequal.

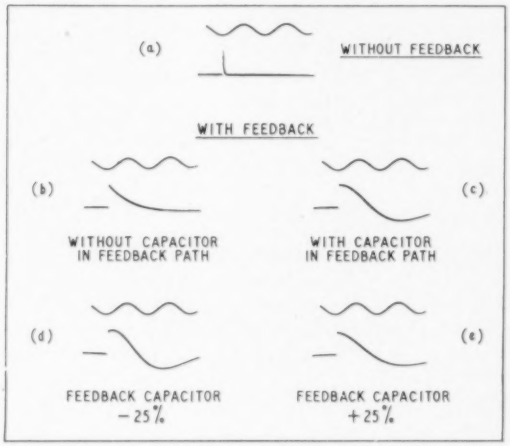


Fig. 4. *Oscillograms of response of model amplifier to a 50-c/s square wave. (The timing wave is 500 c/s.)*

10. Experimental Work

Fig. 3 shows an amplifier which was built in order to obtain some experimental verification of the theory. The values of the components determining the time-constants of the coupling circuits were accurate to within 1%. The time-constants of the decoupling circuits are of the order of 100 times those of the coupling circuits and β-path, which were made much smaller than normal in order to avoid distortion being introduced by the oscilloscope amplifier. The effects of time-constants other than those of the coupling circuits and the β-path are therefore negligible and the analysis given in Section 5.2 is applicable.

The screen voltage of the valves was adjusted so that the voltage gain at high frequencies with the feedback path disconnected was 135, then $(1 + \mu_0\beta_0) = 20$. The value of capacitance required in the feedback path to obtain first-order compensation [calculated from equation (24)] was 0·838 μF.

*i.e., the circuit of Fig. 1 with the feedback path modified as shown in Fig. 3.

Oscillograms were taken of the response of the amplifier to a 50-c/s square-wave input voltage. Because the period of this wave is long compared with the amplifier time-constants, the response to each edge of the waveform is substantially independent of its predecessors and approximates closely to the indicial response. A double-beam oscilloscope was used with the output voltage deflecting one beam and a 500-c/s sinewave calibrating signal deflecting the other beam.

Fig. 4(a) shows the response to the 50-c/s square wave of the amplifier without feedback and Fig. 4(b) shows the response with feedback but without the capacitor in the β-path. Negative feedback has made the response decay much more slowly but the response is still of the zero-order type. Fig. 4(c) shows the response with the 0·838-μF capacitor in the β-path. The initial flatness of the response indicates that first-order compensation has been obtained, but it is accompanied by a large amount of overshoot. Figs. 4(d) and 4(e) show the effect on the response of a $\pm$ 25% variation of the capacitance in the β-path. Oscillograms were also taken of the response of the amplifier to a 500-c/s square wave. Fig. 5 shows the results obtained (a) with β independent of frequency and (b) with the 0·838-μF capacitor in the feedback path.

The gain–frequency response of the amplifier was measured with and without feedback and the results obtained are shown in Fig. 6. Application of negative feedback makes the gain–frequency response flat down to very much lower frequencies, but the inclusion of the capacitor in the feedback path causes a considerable peak in the response, thus confirming that a flat gain–frequency response is not the criterion for a flat indicial response.

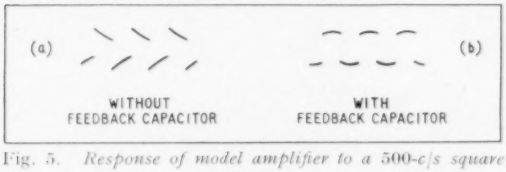


Fig. 5. *Response of model amplifier to a 500-c/s square wave.*

11. Conclusions

The method outlined in Part 1 for designing amplifiers with indicial responses having the highest possible order of compensation is applicable to negative-feedback amplifiers. If the amplifier has a feedback fraction β which is independent of frequency, the overall response has rth-order compensation when the μ-path has an rth-order response. If the μ-path is designed to be maximally flat the overall indicial response will also be maximally flat.

If β is made frequency-dependent, an amplifier can have a higher order of compensation than is obtainable from the μ-path alone. It is shown that the effect on the order of compensation of adding a capacitor to the β-path is equivalent to adding a stage of anode decoupling to the μ-path. This assists in offsetting the effects of the coupling circuits and enables compensation to be obtained which is one order higher than that obtainable when β is independent of frequency.

The method has been applied to some simple circuits, including cases where a coupling circuit or an anode decoupling circuit is outside the feedback loop. Some of the results have been verified by experiment. A high order of compensation of the indicial response does not necessarily imply stability. It is still necessary to check the design of the amplifier to ensure that it is stable.

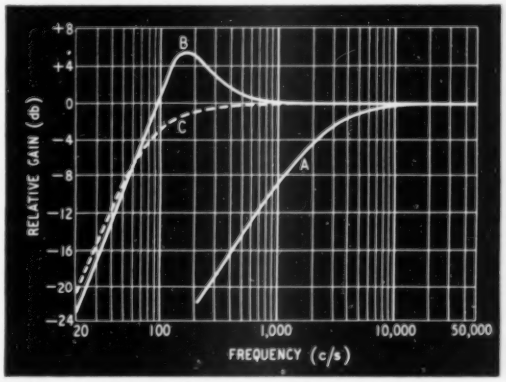


Fig. 6. *Frequency response of a model negative-feedback amplifier; A, without feedback; B and C, with and without a capacitor in the feedback path.*

Acknowledgments

Acknowledgment is made to Siemens Brothers & Co., Ltd., for permission to publish the paper. The author is greatly indebted to Mr. A. G. Simms for his help in the mathematical work contained in the Appendices. Thanks are also due to Mr. H. A. Showell for his assistance in the experimental work.

NATIONAL RADIO EXHIBITION

The 21st National Radio Show will be held at Earl's Court, London, from 25th August to 4th September inclusive (except Sunday). It will be open from 11 a.m. until 10 p.m. and admission will cost 2s. 6d. (children 1s.)

TECHNICAL LITERATURE

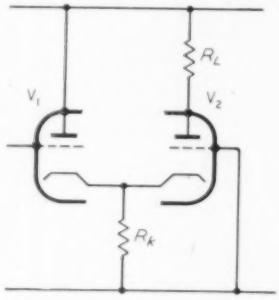
Marconi Instruments 1954 Catalogue

A 200-page bound catalogue which deals with over 70 different equipments for telecommunications measurement and industrial electronics. Functional diagrams of the apparatus are included. Available to senior engineers and executives on request to Marconi Instruments, Ltd., St. Albans, Herts.

CORRESPONDENCE

Cathode-Coupled Amplifier Formulae

SIR,—The discussion of the cathode-coupled amplifier by T. W. Brady[6] in the May issue of *Wireless Engineer* is an excellent addition to the knowledge of this circuit, which presently is finding many applications; for example, in extremely wide band distributed amplifiers (cut-off frequency approaching 1,000 Mc/s)[3,5]. With direct reference to Fig. 1, reprinted from Mr. Brady's article,[6] this writer wishes to add some additional formulae, covering the grid input admittance, the cathode impedance, and the gain. These formulae have been found highly useful, and perhaps provide a complement to Mr. Brady's graphical method.



The input admittance has the form

$$\mathbf{Y}_{in} \approx j\omega C_{gk} \frac{g_{m2} + 1/R_k + j\omega C_k}{g_{m1} + g_{m2} + 1/R_k + j\omega (C_{gk} + C_k)}, \quad (1)$$

where R_k is the cathode resistor, C_{gk} grid–cathode capacitance, and C_k the shunting stray cathode capacitance. Since at higher frequencies the cathode lead inductance L_k plays an unfortunate role in creating a conductance across the tube input, the formula including L_k is of interest (for simplicity given for identical tubes),

$$\mathbf{Y}_{in} \approx j\omega C_{gk} \frac{1 + g_m R_k - \omega^2 L_k C_k + j\omega (g_m L_k + C_k R_k)}{\left[1 + 2g_m R_k - \omega^2 L_k (C_{gk} + C_k) + j\omega [2g_m L_k + R_k (C_{gk} + C_k)]\right]} \quad (2)$$

The cathode output impedance $\mathbf{Z}_0$ seen through the cathode impedance $\mathbf{Z}_k$ looking into the circuit with R_L extended to become the load impedance $\mathbf{Z}_L$, is, for identical tubes,

$$\mathbf{Z}_0 \approx \frac{r_a (r_a + \mathbf{Z}_L)}{(\mu + 1)(2r_a + \mathbf{Z}_L)}, \quad (3)$$

This formula may be easily derived if Thévenin's theorem is applied to the constant-voltage equivalent circuit of the cathode-coupled amplifier, and the writer's 'output impedance theorem' applied.[4]

The complex gain has the form, with R_k extended to become $\mathbf{Z}_k$,

$$\mathbf{A}(\omega) \approx \frac{g_{m1} g_{m2} \mathbf{Z}_L}{g_{m1} + g_{m2} + 1/\mathbf{Z}_k} \quad (4)$$

The use of the 'Korman equivalent triode'[1,2] is often helpful in computation work on the cathode-coupled amplifier (in U.S.A. often referred to as the 'paraphase' amplifier). The tube coefficients for the equivalent triode are

$$\mu_{eq} = -\mu \frac{\mathbf{H}}{1 + \mathbf{H}}, \quad (5)$$

$$r_{aeq} = r_a \frac{1 + 2\mathbf{H}}{1 + \mathbf{H}}, \quad (6)$$

$$g_{meq} = -g_m \frac{\mathbf{H}}{1 + 2\mathbf{H}}, \quad (7)$$

$$\mathbf{H} = g_m \mathbf{Z}_k \left(1 + \frac{1}{\mu}\right) \approx g_m \mathbf{Z}_k. \quad (8)$$

As an example of the use of this method, the approximate gain formula in Equ. (4), for the case of $g_{m1} = g_{m2}$, may be quickly derived, considering the cathode-coupled amplifier as an equivalent triode with a plate load $\mathbf{Z}_L$.

HARRY STOCKMAN.

Scientific Specialties Corporation,
Boston 35, Massachusetts, U.S.A.
27th May 1954.

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STANDARD-FREQUENCY TRANSMISSIONS

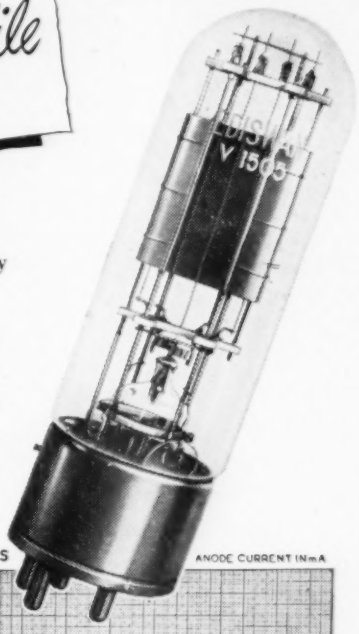
(Communication from the National Physical Laboratory)

Values for June 1954

Date 1954 June	Frequency deviation from nominal: parts in 10^8		Lead of MSF impulses on GBR 1000 G.M.T. time signal in milliseconds
	MSF 60 kc/s 1429-1530 G.M.T.	Droitwich 200 kc/s 1030 G.M.T.	
1	— 1·2	— 1	+ 36·4
2	— 1·1	— 1	+ 36·9
3	— 1·1	— 2	+ 36·5
4	— 1·1	— 1	NM
5	— 1·1	— 2	NM
6	— 1·2	— 2	NM
7	— 1·2	— 1	NM
8	— 1·1	— 1	+ 36·4
9	— 1·1	— 1	+ 36·4
10	— 1·3	— 1	+ 35·4
11	— 1·2	— 2	+ 35·0
12	— 1·1	— 2	NM
13	— 1·1	— 2	NM
14	— 1·2	— 2	+ 34·6
15	NM	— 1	+ 33·6
16	— 1·1	— 1	NM
17	— 1·1	— 1	+ 33·2
18	— 1·1	— 1	+ 33·1
19	NM	— 3	NM
20	NM	— 2	NM
21	— 1·1	0	NM
22	— 1·1	— 1	+ 36·5
23	— 1·1	0	+ 37·0
24	— 1·1	— 1	+ 37·7
25	— 1·1	0	+ 38·4
26	— 1·1	— 5	NM
27	— 1·1	— 1	NM
28	— 1·1	0	+ 40·3
29	— 1·0	0	+ 40·8
30	— 1·0	— 1	+ 41·1

The values are based on astronomical data available on 1st July 1954.
NM = Not Measured.

DATA for the Designers File



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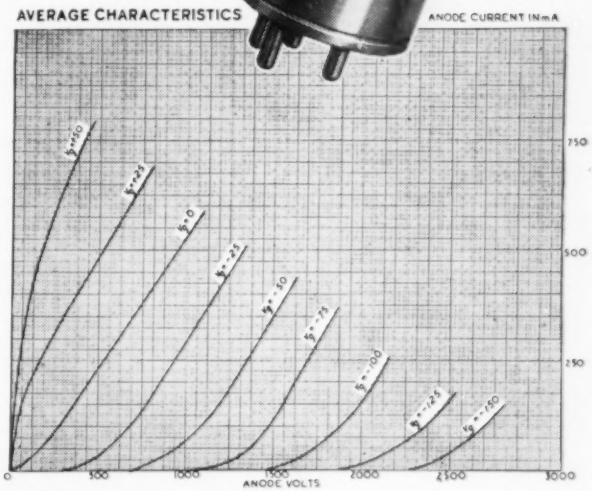
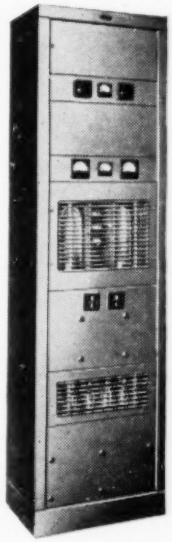
Amplification Factor	16*
Anode Impedance (r_a)	2,000 ohms*
Maximum Operating Frequency at full rating	1.5 Mc/s
Audio Power Output	1,000 watts (2 valves)

* *Taken at V_a = 2,000 volts; I_a = 150 m/A.*





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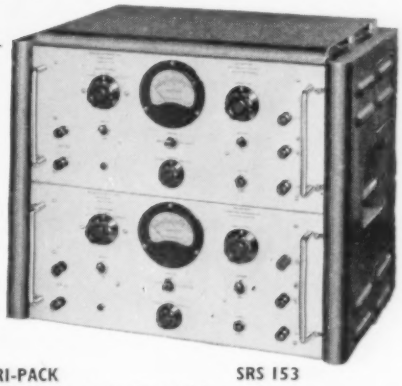

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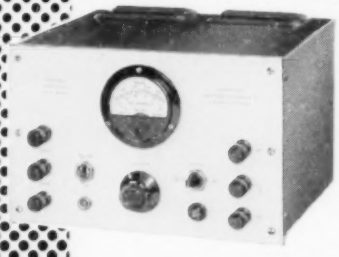
Ripple: Less than 8 millivolts.
A.C. Output: Two 6·3 volt outputs at 5 amps.



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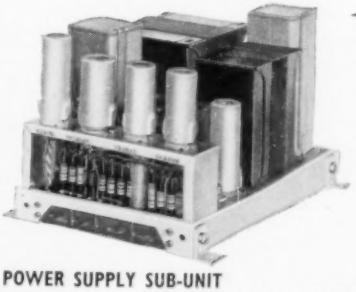

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D.C. Output Current: 0—50 m/a.
A.C. Output: 6·3 volts 4 amps. C.T. 6·3 volts 1 amp.
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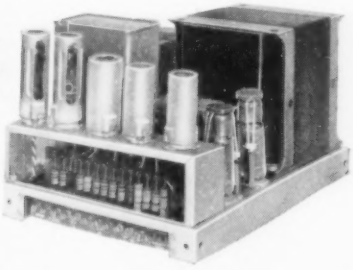
A.C. Impedance: 0·05 ohm.
Ripple and Noise Content: Less than 200μ volts.



POWER SUPPLY SUB-UNIT **SRS 155B**

D.C. Output Voltage (Isolated): 250 volts + Ve or − Ve.
D.C. Output Current: 10—100 m/a.
A.C. Output: 6·3 volts 4 amps C.T. 6·3 volts 2 amps C.T. 6·3 volts 1 amp.
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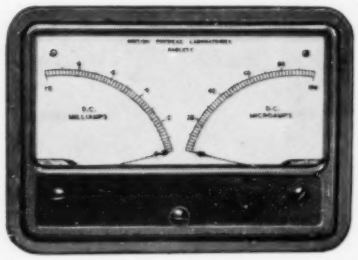

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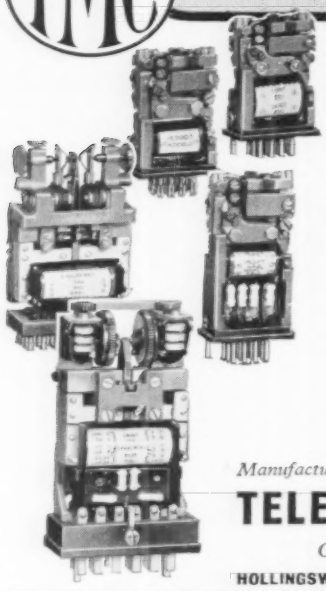

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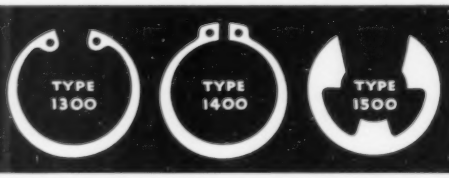

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(*a*) The general engineering development of receiving valves.

(*b*) Glass techniques, including the control of existing and experimental work on new methods.

(*c*) The investigation of manufacturing processes with a view to attaining closer control and further development.

REF: W.2 **Gas-Filled Valve Development Department**
(*a*) The technological development of hot cathode gas-filled rectifiers and thyratrons of all sizes and classes.

(*b*) The general development of cold cathode gas-filled valves including stabilisers and reference tubes and also multi-electrode types.

(*c*) Investigational work on gas discharge phenomena in relation to the tube classes mentioned under (*a*) and (*b*) above.

REF: W.3 **U.H.F. and Transmitting Valve Development Department**
(*a*) The technological development of magnetrons, klystrons, travelling wave tubes and kindred devices. This work includes metal-glass sealing, ceramic and metal brazing techniques.

(*b*) General development work on the more conventional glass, metal-glass and disc seal valves usually for the V.H.F. range.

(*c*) Investigational work on the effects of variation of mechanical structure and of changes in materials used for (*a*) and (*b*) above.

REF: W.4 **Cathode Ray Tube Department**
(*a*) The glass, mechanical and chemical technology of all types of cathode ray and kindred tubes including colour television tubes.

(*b*) Process development of methods used for the manufacture of new types of cathode ray tubes.

(*c*) Improvements in gun design including electron optical work.

REF: W.5 **Semi-conductor Development Department**
(*a*) The technological development of all types of semi-conducting devices including transistors and rectifiers.

(*b*) The investigation of semi-conducting materials to establish their possible application to practical devices.

(*c*) The development of processes and production methods for semi-conducting devices.

Applicants for the senior posts should be physicists, electrical engineers, or, where appropriate, physical chemists and should possess a university degree or equivalent qualifications.

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Assistants are required for the following vacancies:—

REF: W.6 The design and construction of apparatus and equipment for testing radio valves and other vacuum devices. A knowledge of pulse techniques is involved.

REF: W.7 The testing by Quality Control methods of Cathode Ray Tubes of all types. Liaison work with other departments will be required.

REF: W.8 The supervision of a small section carrying out quality control testing and special measurements of receiving valves. Liaison work with other departments will be required.

REF: W.9 The testing of klystrons and similar U.H.F. devices and the development of testing techniques for this purpose.

REF: W.10 Testing and experimental work on semi-conducting devices of the crystal diode and transistor type.

Applicants for the above posts, which are permanent, should be Physicists or Electrical Engineers and should possess a university degree or equivalent qualification.

In addition to the above posts there are a number of less senior posts available for work in similar fields, for applicants who possess Advanced Level G.C.E. in science subjects or equivalent qualifications, or who are pursuing courses leading to H.N.C. or external degrees. Such applicants should quote reference W.17.

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REF: W.11 (1) Gas or mercury vapour thyratrons and rectifiers.

REF: W.12 (2) Ultra high frequence valves such as magnetrons, klystrons and travelling wave tubes.

REF: W.13 (3) Transmitting valves for communications and general industrial uses.

REF: W.14 (4) Pulse techniques associated with U.H.F. devices and thyratrons.

REF: W.15 (5) Receiving valves used for frequencies between 100 and 1,000/Mc/s.

REF: W.16 (6) Valves and semi-conductors used in computing and switching apparatus.

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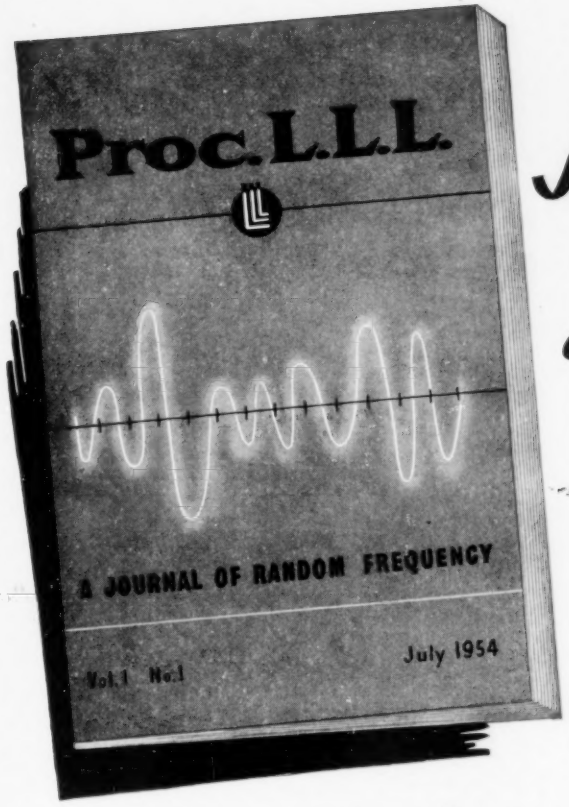

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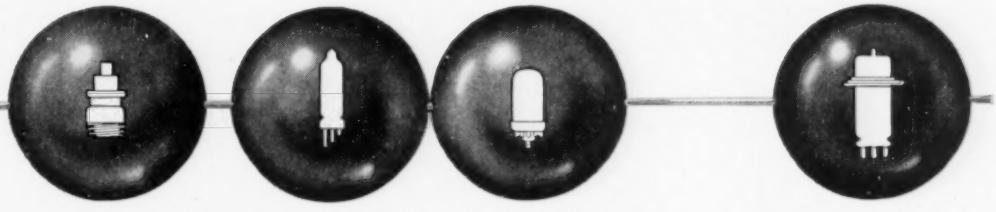

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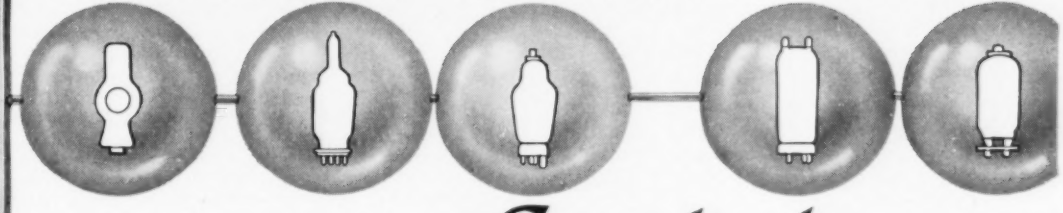

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B.B.C. requires Engineer for Cinefilm Equipment Unit, in London. Applicants should be qualified engineers, preferably with degree in electrical engineering or physics and should have had considerable practical experience in use of modern high-grade cinematographic equipment, thorough knowledge of the principles of its design and of associated techniques relating to lenses, and of application of films for television purposes. First-class technical knowledge of such equipment and practice, including use of picture and sound-recording cameras, projectors, film stock, film processing, etc., and good knowledge of principles of television are called for. Experience in direction of other engineers desirable. Duties include planning, specification, ordering and testing of such equipment and conduct of correspondence. Starting salary £990 rising by annual increments to £1,320. Apply: E.E.O., B.B.C., London, W.1, quoting ref. 'E.943.W.E.'

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Please write giving details to the Personnel Officer, Ericsson Telephones, Ltd., Beeston, Nottingham. (Reference LK/3).

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B.B.C. requires Engineer for Cinefilm Equipment Unit in London. Applicants should have University degree in electrical engineering or physics or equivalent qualification and good technical knowledge of modern cinematographic practice, including picture and sound-recording cameras, projectors, film stock, film processing, etc. Duties include planning, specification, ordering and testing of equipment and conduct of correspondence. Starting salary £645 rising by five annual increments to £880. Apply: E.E.O., B.B.C., London, W.1, quoting ref. 'E.940.W.E.'

B.B.C. requires Engineer for Television Unit, Studio Equipment Section, in London. Applicants should have University degree in electrical engineering or physics or equivalent qualification and knowledge of principles of design of high-grade cinematographic equipment, preferably with practical experience in use of film in relation to television, and good knowledge of principles of television. Duties include specification, ordering, testing of equipment used for transmission of and making films (telerecordings) from television pictures, and under supervision planning, estimating and execution of projects involving this type of equipment, and liaison with manufacturers. Starting salary £645 rising by five annual increments to £880. Apply: E.E.O., B.B.C., London, W.1, quoting ref. 'E.941.W.E.'

B.B.C. requires Engineer with University degree or equivalent, in electrical engineering, for Transmitter Section of Planning and Installation Dept., in London. Specialised knowledge of transmitter drive technique as applied to precision crystal and variable frequency checking equipment desirable. Knowledge of V.H.F. and S.H.F. transmitting and receiving equipment and television principles an advantage. Duties include specification, installation and acceptance lists of such equipment. Applicants should be prepared to travel to all parts of U.K. as required. Salary £645 (may be higher if qualifications exceptional) rising by five annual increments to £880 max. Promotion prospects. Requests for application forms quoting ref. E.928.W.E. should reach E.E.O., Broadcasting House, W.1, within five days.

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Apply, giving full details of qualifications, experience, etc., to the Personnel Dept. (RL/10), E.M.I., Ltd., Hayes, Middx.

An opportunity occurs in progressive manufacturing concern in London area for Engineer aged 25/35 with electrical or physicist degree, on development and research work on high frequency telecommunication cable and associated testing equipment. Salary will be commensurate with qualifications and experience. Apply giving full particulars to Box 6571, c/o *Wireless Engineer*.

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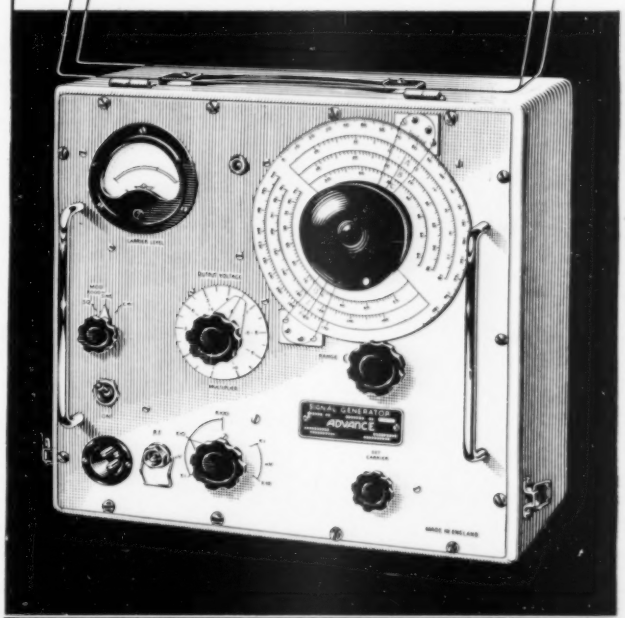

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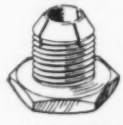

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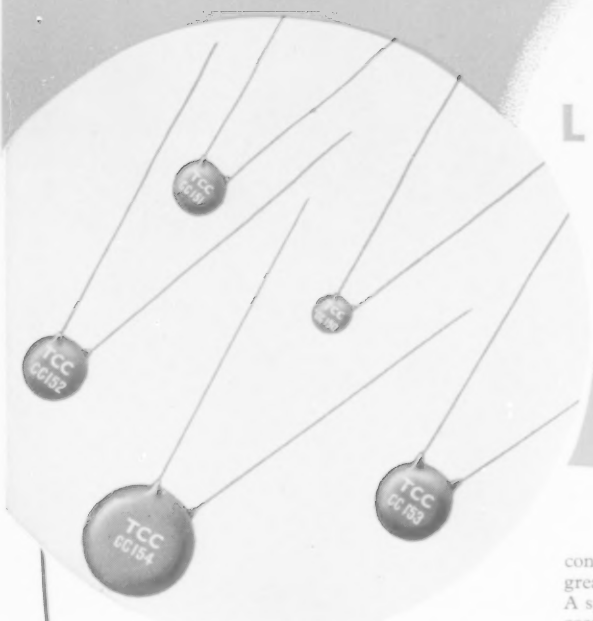

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EFFECT OF TEMPERATURE

In common with other Hi-K ceramic condensers, variations in capacity due to temperature changes may be quite substantial, a fact that should be borne in mind when selecting this type of condenser, and when specifying the capacity required for a particular application. Where such fluctuations of capacity are of no great importance, these Discs may be used as an alternative to condensers using other dielectrics.

CONSTRUCTION

The Dielectric is a specially selected material of the barium-titanate group with a dielectric constant of 3,000, and an insulation resistance greater than 5,000 megohms.

A special non-cracking, heat resistant, protective coating (Finish 'G') is the standard finish. This thermo-setting compound adequately prevents the ingress of moisture, and does not soften or crack at temperatures up to 100°C.; it also provides electrical insulation, thus enabling the condenser to be mounted in contact with the chassis or other components.

Maximum Working Voltage
500v. D.C. or 250v. A.C.

Tolerance of Capacity —20% +80%

Capacity in pF	Dimensions in millimetres		T.C.C. Type Number
	Diameter	Thickness	
1,000	7·5	0·6	CC.150
1,500	7·5	0·6	CC.150
2,000	10·0	0·6	CC.151
3,000	12·5	0·6	CC.152
5,000	15·0	0·6	CC.153
10,000	22·0	0·8	CC.154

Approximately 2mm. should be added to these dimensions to allow for the protective coating.

Terminations are tinned copper wires 50mm. long.

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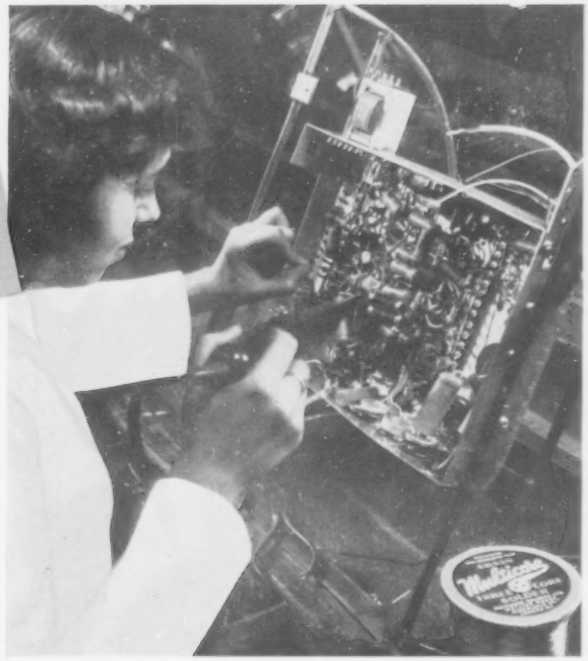

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